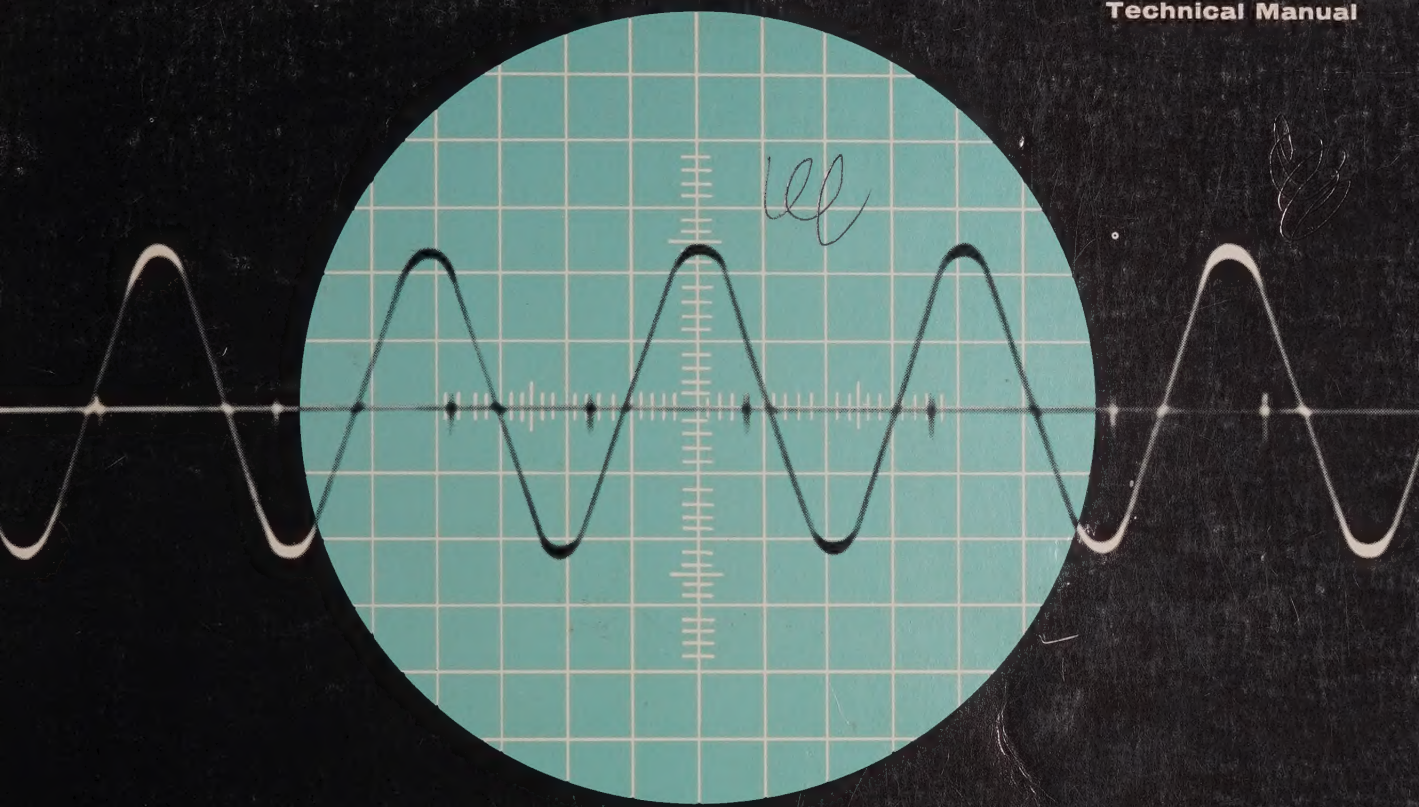
Technical Manual



VARIABLE-RATE, PRECISION

R.F. Sweep Frequency Generator

MODEL 707-D

JERROLD ELECTRONICS CORPORATION
GOVERNMENT AND INDUSTRIAL DIVISION
PHILADELPHIA, PENNA. 19105

7935-572

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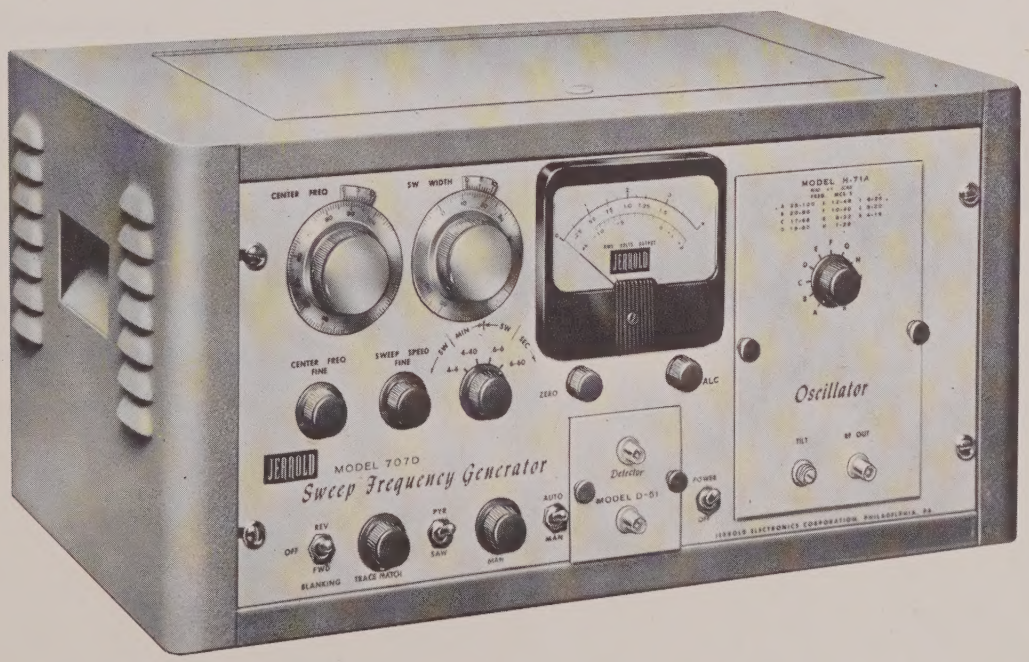
INSTRUCTION MANUAL
435-399.1

VARIABLE-RATE, PRECISION RF SWEEP FREQUENCY GENERATOR

GENERATOR MODEL 707D
PLUG-IN OSCILLATOR HEAD MODEL H-71A
PLUG-IN OSCILLATOR HEAD MODEL H-72A
PLUG-IN OSCILLATOR HEAD MODEL H-73A



JERROLD ELECTRONICS CORPORATION
GOVERNMENT AND INDUSTRIAL DIVISION
PHILADELPHIA, PA. 19132

LKE, January 1966

NOTE: This manual is shipped with:
Power Supply Model 707-D Serial # _____

This manual incorporates all Engineering Change Orders up to and including E. C. O. _____

All data in this manual are subject to change without notice.

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Replacement Parts Lists	7-1 to 7-10
SCHEMATIC CIRCUIT DIAGRAMS	DWG. NO.
Power Supply Model 707-D	E861-538
Plug-in Oscillator Model H-71A	D861-277
Plug-in Oscillator Model H-72A	D861-368
Plug-in Oscillator Model H-73A	D861-367
Plug-in Detector Model D-51	861-366

RF SWEEP FREQUENCY GENERATOR MODEL 707-D

INTRODUCTION

Jerrold Model 707-D is an instrument especially designed to investigate the frequency response of amplifiers, networks, and components within a frequency range from 2 to 200 mc/s, depending on the type of plug-in oscillator head used (see specifications for complete data).

The instrument features an extremely flat output, with harmonics 40 db down over the highest octave, 30 db down over the full range. Sweep rates are variable from 60 sweeps per second to one sweep every 2-1/2 minutes with a sweep width variable from $\pm$ 1/2% to $\pm$ 60% of center frequency. The instrument is thus highly suitable for use with either an X-Y plotter or an oscilloscope having a long-persistance cathode ray tube.

Special switching circuitry permits either manual or automatic sweeping, producing sweep voltage waveforms of either pyramidal or sawtooth characteristics.

Other features include: no blanking, or blanking of either forward or return trace; means of insuring that forward and return traces are in phase; and a metered r.f. output with the meter having both voltage and dbm scales.

The instrument can be energized from either a 115 vac 50 to 60 cps, or a 230 vac 50 to 60 cps source.

The unit is supplied in a bench type portable cabinet. Front panel dimensions are such that, without the cabinet, the unit can easily be mounted on a 19" standard relay rack.

SPECIFICATIONS

1. SWEEP GENERATOR MODEL 707-D

SWEEP RATE	From 60 sweeps per second to one sweep every 2 1/2 minutes
SWEEP METHOD	Automatic or manual
SWEEP VOLTAGE WAVE FORM	Pyramidal (50:50 rise time to fall ratio) or sawtooth (Approx. 80:20 rise time to fall ratio)
BLANKING	No blanking, or blanking of either forward or return trace
OUTPUT METER	red scale 0 to 1.25 volts black scale 0 to 4 volts dbm scale -7 to +5 dbm
PLUG-IN DETECTOR MODEL D-51B (included in basic unit)	Full wave, peak-to-peak, voltage doubler. VSWR less than 1:1; maximum voltage input 3 volts rms; flatness $\pm$ 1/2 db to 220 mc/s; insertion loss 1 db; impedance 50 ohms - BNC connectors
TUBE COMPLEMENT	(3) OD3, (2) 12AT7, (2) 6U8, (1) 6K6, (1) 12BY7A, (2) 6AU6, (1) 6CL6, (1) 12AX7
POWER INPUT	115 VAC, 50 - 60 CPS: 230 VAC, 50 - 60 CPS
POWER CONSUMPTION	100 watts

2. PLUG-IN OSCILLATOR HEADS

	H-71A	H-72A	H-73A
FREQUENCY RANGE	10 to 100 mc/s in positions A to K	16 to 200 mc/s in positions A to K	2 to 50 mc/s in positions A to K
OUTPUT IMPEDANCE	50 ohms for all oscillator heads		
OUTPUT VOLTAGE	20 dbm 3 volts rms into a 50 ohm load on all ranges	13 dbm 1 volt rms into a 50 ohm load on all ranges	20 dbm 3 volts rms into a 50 ohm load on all ranges
OUTPUT VOLTAGE VARIATION	+ 0.05 db from 50 to 100 mc/s + 0.15 db over full range	+ 0.05 db from 100 to 200 mc/s + 0.15 db over full range	+ 0.05 db from 25 to 50 mc/s, ± 0.1 db over full range
HARMONICS	40 db down over highest octave, 30 db down over full range; all model		
SWEEP WIDTH	Variable from ± 1/2% to ± 60% of center frequency; all models		
LINEARITY	The instantaneous frequency corresponds to the instantaneous horizontal deflection voltage within ± 2% of the peak-to-peak voltage at a 2:1 sweep width; all models		
CONNECTORS	50 ohm BNC for all models		
TUBE COMPLEMENT	(2) 12BY7A	(1) 6922	(2) 12BY7A

CHART OF CONTROLS AND CONNECTIONS

FRONT PANEL			
NAME AND POSITION	CIRCUIT DESIGNATION	TYPE	FUNCTION
POWER - OFF	S1	2 POSITION TOGGLE SWITCH	Energizes and de-energizes the unit.
CENTER FREQ	R44	POT.	Sets center freq. of sweep.*
CENTER FREQ FINE	R45	VERNIER POT.	Fine adjustment of center freq. as set by R44.
SW WIDTH	R40	POT.	Controls frequency deviation of sweep. (see specs.)
SW/MIN - SW/SEC .4 - 4 4 - 40 .6 - 6 6 - 60	S5	4-POSITION WAFER SWITCH	Selects range of sweep rate. 0.4 to 4 sweeps per minute. 4 to 40 sweeps per minute. 0.6 to 6 sweeps per second. 6 to 60 sweeps per second.
SWEEP SPEED FINE	R31	VERNIER POT.	Selects sweep rate within range set by S5.
BLANKING FWD OFF REV	S2	3-POSITION TOGGLE SWITCH	Controls blanking mode of trace. Blanks forward trace. No blanking. Blanks reverse trace.
TRACE MATCH	R41	POT.	Permits phase adjustment (with S2 in OFF position) for trace and retrace waveform on scope.
PYR SAW	S3	2-POSITION TOGGLE SWITCH	Provides pyramidal output. Provides sawtooth output.
AUTO MAN	S4A, B	2-POSITION TOGGLE SWITCH	Sweeps range automatically. Permits manual sweeping by R25.

* Dial knob graduations are for reference only and do not indicate frequencies.

CHART OF CONTROLS AND CONNECTIONS

FRONT PANEL (Cont)			
NAME AND POSITION	CIRCUIT DESIGNATION	TYPE	FUNCTION
MAN	R25	POT.	Manual sweeping.
ALC	R52A, B	DUAL-POT.	Adjust automatic level control and r.f. output circuit.
ZERO	R56	POT.	Permits zeroing meter when oscillator is off.
DETECTOR MODEL D-51B		PLUG-IN	Unit contains detector circuit bridged across a transmission line having the impedance for which the sweep is designed. Either connector, J4 or J5, is usually terminated by a Jerrold Model TR-50B.
CHASSIS TOP			
SWING	R6	POT.	Controls amplitude of pyramid.
DC REF	R9	POT.	Controls level of pyramid with respect to ground.
SYM	R27	POT.	Controls symmetry of pyramid.
LIN A LIN B	R38 R35	POT. POT.	Adjust linearity of sweep vs. frequency.
METER SET	R55	POT.	Calibrates meter M1.

CHART OF CONTROLS AND CONNECTIONS

CHASSIS REAR			
NAME AND POSITION	CIRCUIT DESIGNATION	TYPE	FUNCTION
SWEEP SAMPLE OUTPUT	J3	BNC CHASSIS FITTING	Connection of sweep output to marker mixer.
BIRDIE AUDIO INPUT	J8	BNC CHASSIS FITTING	Connection of audio input from marker mixer.
VERT	J7	BNC CHASSIS FITTING	Connection of detector output to vertical input of scope.
HORIZ	J6	BNC CHASSIS FITTING	Connection of horizontal deflection voltage to horizontal input of scope.
LINE SYNC ON-OFF	S6	2-POSITION TOGGLE SWITCH	In ON position, locks sweep rate to 60 cycle line for operation with Model FD-30 Coaxial Switch.
SLO-BLO	F1	FUSE	2A-115 V/1A - 230 V; line protection.
SWEEP DRIVE OUT	J1	6-PIN SOCKET	Connection to X-Y plotter, etc.

CHART OF CONTROLS AND CONNECTIONS

PLUG-IN OSCILLATOR — MODELS H-71A, H-72A, H-73A

FRONT PANEL			
NAME AND POSITION	CIRCUIT DESIGNATION	TYPE	FUNCTION
A thru K	H-71A:S1 H-72A:S201 H-73A:S301	11-POSITION ROTARY SWITCH	Chooses one of the 11 frequency ranges noted above switch.
TILT	H-71A:C28 H-72A:C224 H-73A:C329	h.f. TRIMMERS	Permits about 0.25 db of tilt adjustment.
R. F. OUT	H-71A:J1 H-72A:J201 H-73A:J301	BNC CHASSIS FITTING	R. F. output cable connection.
CHASSIS			
BAL	H-71A:C36, 37 H-72A:C214, 217 H-73A:C317, 320	VARIABLE CAPACITORS	For balancing push-pull output circuit, to obtain minimum harmonics.
HARMONIC BALANCE	H-71A:R2 H-72A:R202 H-73A:R302	POT.	
METER INPUT	H-71A:R11 H-72A:R212 H-73A:R311	POT.	Adjusts dc input voltage to meter M1.
NOTE: These four chassis-mounted controls are factory-aligned; adjustment may be needed only after oscillator tube replacement.			

CIRCUIT DESCRIPTION

(See Dwg. E861-538)

1. POWER SUPPLY CIRCUIT.

AC power at the input is fused in F1 with 2 amps. for 115 vac 50 - 60 cps, and with 1 amp. for 230 vac 50 - 60 cps. Terminal board TB1 permits strapping the line transformer T1 for either ac source. T1 is a multiple winding transformer providing three separate 6.3 vac outputs for tube heaters, and from a fourth secondary winding, through silicon rectifiers CR1, CR2 and CR3, the necessary positive and negative dc voltages. Four such voltages are derived, three of which are regulated by V1, V2 and V3. Switch S1 connects the unit to and disconnects it from the ac line.

2. SWEEP CIRCUIT.

The basic sweep frequency circuit employs tubes V4, V5 and V6 with their associated components. The two triodes of V4 constitute a simple flip-flop circuit producing a pulse with a rectangular wave shape. The pulse amplitude is controlled by SWING control R6. A chassis-mounted DC REF potentiometer R9 controls the output level of the pulse with respect to ground. This pulse is first fed to tube V5, the triodes of which are arranged in a cathode follower circuit. From there, the pulse is passed via switch S3 to an integrating circuit comprising SWEEP SPEED FINE potentiometer R31 and one of the capacitors C7, C8, C9 or C10, as chosen by the SW/MIN-SW/SEC switch S5. With switch S3, the pulse can be chosen to exhibit either a pyramidal or a sawtooth waveform. With S3 in PYR position, the symmetry of the pyramidal pulse is controlled by SYM potentiometer R27. With S3 in SAW position, the pulse is made unsymmetrical with respect to time and the return rate of the sweep is faster than the forward rate; integration of this pulse produces a sawtooth waveform. For increasing the linearity of the time-base voltage, tube V6 is employed as a "Miller" high-gain amplifier in parallel to the integrating rc network.

From the cathode of V-6B the pulse is then passed via SW WIDTH potentiometer R40 and a voltage divider network to the control grids of V7 and V8. The function of these two tubes is not only to drive the reactor coil in the plug-in oscillator, but by driving it from two tubes with complementary characteristics, the waveform of the current flowing through the coil is slightly changed so that maximum linearity with respect to frequency is achieved on the scope display of the sweep. LIN A potentiometer R38, and LIN B potentiometer R35, permit adjusting the characteristics of V7 and V8 for best linearity.

TRACE MATCH potentiometer R41 insures that, with switch S2 in OFF position (no blanking), the r.f. output observed on the oscilloscope will be in phase during trace and retrace.

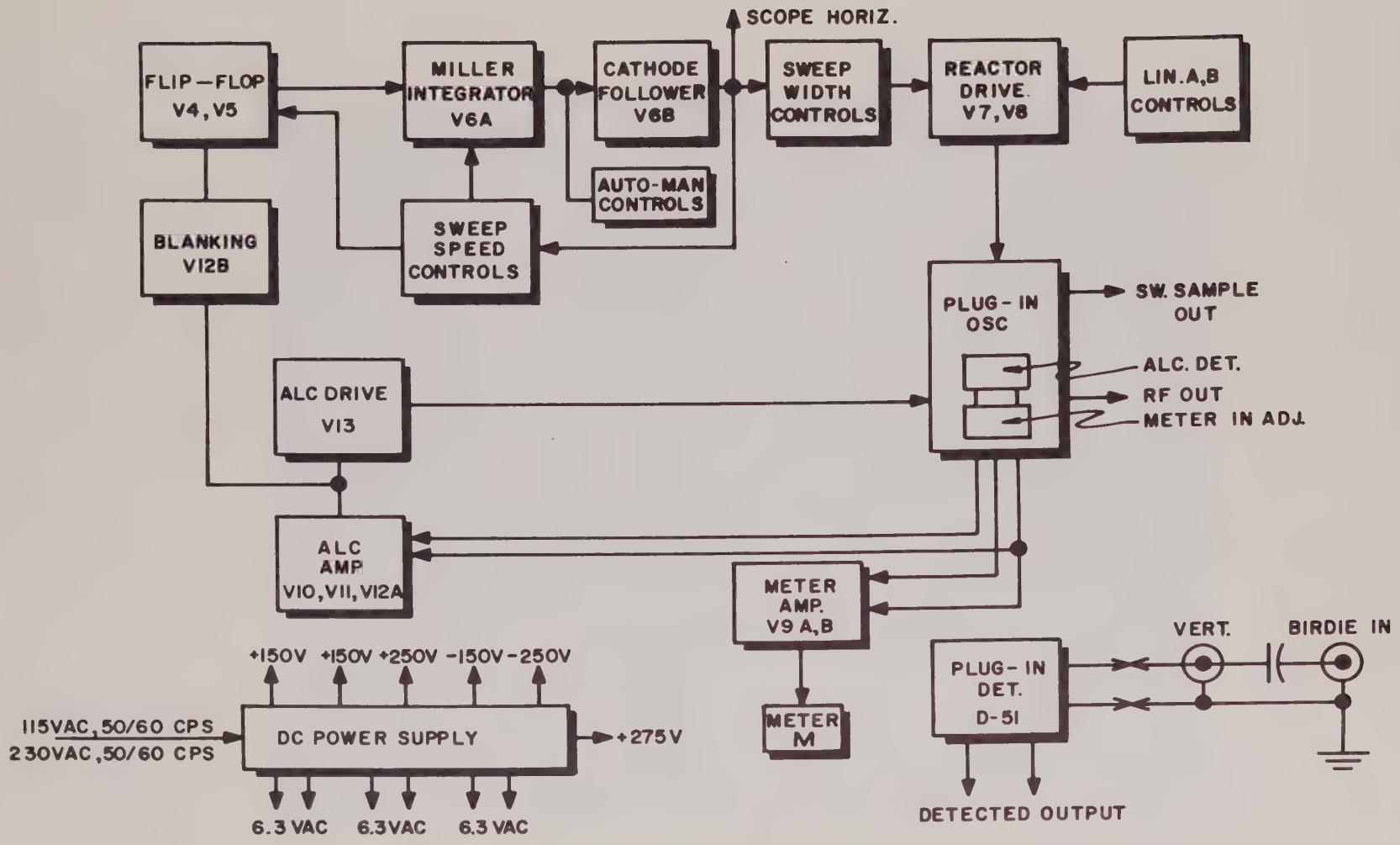
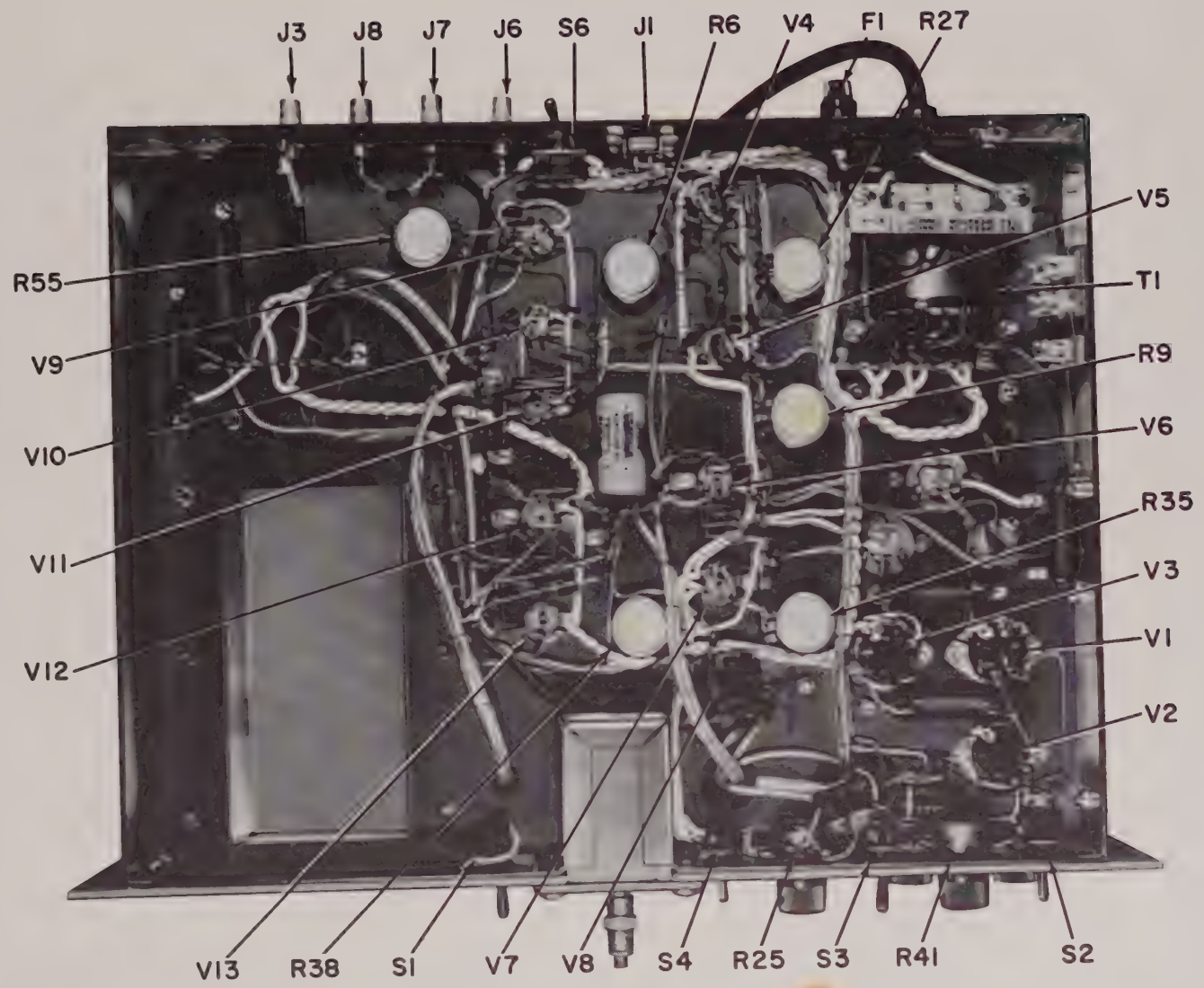


Figure 1. Model 707-D, Functional Block Diagram



Component Location Diagram

AUTO MAN switch S4 permits either automatic or manual sweeping; fine control of automatic sweep speed is exercised by potentiometer R31 in the integrator circuit. Manual sweeping is controlled by MAN potentiometer R25. Two potentiometers in series, R44 and R45, permit coarse and fine adjustment respectively of the center frequency setting by varying the control grid bias on V7 and V8. Fine adjustment is especially useful when operating at narrow sweep widths.

The plate output of V7 and V8 is then passed via the multi-pin socket J2 to the plug-in oscillator.

3. AUTOMATIC LEVEL CONTROL CIRCUIT.

While the output of the r.f. oscillator is swept through a desired band of frequencies, the level of that output tends to vary. This caused by a number of factors such as the change in Q of the oscillator transformer T1 primary, the extent to which the output coupling varies with frequency, and other related factors.

In order to keep the output voltage constant with frequency, a direct-coupled ALC amplifier circuit is employed, using tubes V10 through V13 and associated components, whereby the dc voltage from the measuring circuit in the oscillator is applied to the grids of the first tube V10. The amplitude of this dc voltage is a measure of the r.f. voltage at the tap.

The ALC amplifier circuit terminates in V13 which is connected as a cathode follower. Its output provides the screen grid voltage for the r.f. oscillator. Since the ALC amplifier output voltage is phased for inverse feedback, its level will be extremely constant. Initial level adjustment is achieved by the front panel-mounted ALC dual potentiometer R52 acting as a delay bias control.

The sweep generator is designed to have a characteristic output impedance of 50 or 75 ohms, looking into the r.f. output jack in the oscillator circuit. This is accomplished in the following manner: the output control circuitry keeps the r.f. voltage constant at the junction of the output of the low-pass filter and the 50-ohm, or 75-ohm, resistor (R7 in H-71A head, R208 in H-72A head, or R308 in H-73A head) in the oscillator circuit. For any change in r.f. voltage, the voltage at this junction remains constant, thus the impedance from this point to ground must be zero. Hence, looking from this junction to the r.f. output spigot, one sees a constant voltage source in series with a 50-ohm, or 75-ohm resistance, as the case may be.

4. METERING.

The metering circuit employs a twin-triode, tube V9A and B. The grids of the triodes are interconnected through resistor R58. With no voltage impressed across this resistor, i.e., when the oscillator has zero output, the front panel-mounted ZERO potentiometer R56 can be adjusted so that no current flows through the meter M1. With a dc voltage impressed across R58, tube V9 experiences a change in conduction, triode V9A

increasing and V9B decreasing. This change produces a difference in potential across the meter. The circuit also incorporates METER SET potentiometer R55 for meter calibration. R55 is accessible from the chassis top for screwdriver adjustment.

5. BLANKING.

It is often convenient to be able to turn the sweep generator off periodically in order to establish a zero output reference level for a circuit under investigation. For this purpose, Model 707-D has a built-in blanking circuit. Sweeping from one frequency to another is effected by a pyramidal or sawtooth wave obtained by integrating a rectangular pulse. During the first portion of the pulse the generator sweeps in one direction and during the second portion in the opposite direction. Blanking during either portion of the pulse is obtained by simply flipping the BLANKING switch S2 to the desired position. Thus zero level is obtained for either forward or return trace.

The blanking circuit works in the following manner:

Samples of the pulses referred to above are taken from the cathode followers V5A-V5B and applied to two terminals on switch S2. From the common of S2 the pulse sample is fed via resistor R70 to the grid of tube V12B in the ALC amplifier. The plate output of V12B is connected through R72 to the plate of V12A. When S2 is in the OFF position, the grid of V12B will conduct; this essentially puts a short across V12A, hence cuts off the screen grid voltage from the oscillator tube.

6. PLUG-IN OSCILLATOR HEADS H-71A, H-72A AND H-73A

(See Dwgs. D861-277, D861-368 and D861-367)

The circuitry of all three oscillator heads is essentially the same. The heads differ only in their frequency ranges and the values of some of their components.

For simplicity, the circuit description from here on will refer to the schematic designation of components on oscillator head H-71A only.

The oscillator heads are plug-in type sweep oscillators, each with its frequency range subdivided into eleven overlapping ranges as denoted on the oscillator front panel. A rotary switch S1 permits selecting the desired sweep range.

The oscillator circuitry comprises transformer T-1, saturating coil L-12, two pentode tubes V-1 and V-2 operating in push-pull, a splitting filter network, and a voltage measuring circuit employing diodes CR1 and CR2. The tuned circuit essentially comprises T1 with a center-tapped primary winding, stray capacities and one of capacitors C1 through C10 as selected by the rotary switch S1.

The control grids of V-1 and V-2 are connected to the grid winding on the secondary side of T-1 and are phased to permit oscillation. Bias for the control grid of each tube is furnished by grid leak resistors R-1 and R-3, respectively. B+ voltage for the plates of V-1 and V-2 is derived from the center tap of the primary on T-1. Screen grid voltage for both tubes is obtained from the output of the ALC amplifier as previously described. Resistor R-2 is a cathode bias common to both V-1 and V-2; it is variable so that a harmonic balance in oscillations can be achieved.

The oscillator frequency is primarily determined by the capacitor chosen through rotary switch S-1 and the inductance of the T-1 primary as varied by the degree of saturation of the iron core of the transformer. The magnetic flux saturating the iron core is produced by inductor coil L-12 wound on a U-shaped laminated iron core. The currents flowing through L-12 are derived from the sweep circuit and provide the necessary magnetic flux to vary the oscillator frequency. The degree of frequency variation is determined by the variations in the current flowing through L-12.

Output power is taken through a small coupling coil on T-1. One end of this coil is grounded, the other passes the oscillator output to the center of a high-pass low-pass splitting filter. The oscillator output then passes through the low-pass filter section to a junction from where it proceeds: (1) through an RC block to the voltage measuring circuit; (2) through resistor R-7 to the RF OUT jack J-1.

The excellent flatness characteristics of the plug-in oscillator are the result of decreasing the harmonic content of the output, and proper automatic level control action. Second harmonics are reduced through the push-pull circuit configuration and the proper balancing of both tubes by the variables R-2, C36 and C37. Third and higher harmonics are reduced by the splitting filter network.

7. VOLTAGE MEASURING CIRCUIT.

The voltage measuring circuit comprises a voltage divider network and two silicon diodes CR1 and CR2. The network reduces the voltage to prevent overdriving the diodes. CR1 and CR2 act as a voltage doubler rectifier. The amplitude of this dc voltage is a measure of the r.f. voltage at the RF OUT jack J-1.

The output of the diodes is passed via the multi-pin connector J-2 to the ALC and metering circuit in the sweep generator.

A TILT control, high-frequency trimmer C-28, provides tilt adjustment by approximately 0.25 db. Tilt adjustment is made whenever it is necessary to compensate for tilt variations encountered while switching from one oscillator range to another, or whenever attenuators, pads or external detectors are used. Tilt

compensation will then give scope presentations with maximum flatness of the detected sweep output. The control is mounted on the oscillator front panel and adjustable by a screwdriver.

8. PLUG-IN DETECTOR MODEL D-51.

(See Dwg. 861-366)

Model 707-D is equipped with a plug-in type detector, necessary wherever response measurements are to be made of passive networks or preamplifiers, for delay line impedance matching, for checking the sweep output and where such equipment under test does not have a built-in detector.

The detector is designed to have characteristics approaching as closely as possible those of the voltage measuring circuit in the oscillator head. This is essential for presenting a flat output on the oscilloscope, especially at the higher frequencies.

The detector uses diodes CR-101 and CR-102 in a full-wave rectifier circuit which discriminates against even-order harmonics and provides twice the dc output of a half-wave rectifier. The filter network chosen gives a good compromise between carrier filtering and audio response.

OTHER APPLICATIONS

OPERATION WITH X-Y PLOTTER:

When operating the Model 707-D in conjunction with an X-Y plotter, it is desirable to have manual control of the sweep frequency range. This facility is provided by setting switch S4 to MAN position, then by varying R25 marked MAN on front panel, the desired frequency range can be swept manually. The circuitry is so arranged that if the sweep is set manually at any frequency range and then switched to automatic sweep, the generator will start sweeping at the preset frequency. This is insured by R24 marked MAN SET on the chassis; R24 is a factory adjusted control.

OPERATION WITH MODEL FD-30 VOLTAGE COMPARATOR:

Often it may be desired to use Model 707-D in conjunction with Jerrold Model FD-30 Wide-Band Comparator.

To operate the Model 707-D Generator effectively with the Model FD-30 Voltage Comparator requires that the sweep rate of the Generator can be phased with the switching rate of the Voltage Comparator. Since the switch rate of the Voltage Comparator is locked to the 60 cycle line, this requires that the sweep rate of the Generator also be locked to the line when the two units are to be operated together. A simple method of accomplishing this is to feed a small amount of line voltage to the grid of the Model 707-D flip-flop circuit for the purpose of locking the sweep rate circuit to line when the time constants of the integrating circuit are set for a 60 cycle rate.

To make this possible, Model 707-D has a special LINE SYNC switch (S6) mounted at the rear of the chassis. This switch locks the generator's sweep rate to the 60-cycle line at either a 60-cycle, 30-cycle or 15-cycle rate.

EQUIPMENT MAINTENANCE

GENERAL

Model 707-D has been designed to give trouble-free long-life service. Components incorporated have been carefully selected to insure a high measure of reliability and accuracy. No major trouble should therefore be encountered while the instrument is used by qualified personnel and not subjected to abusive handling.

Should it become necessary to replace a tube, only a new, exactly equivalent type of a recognized brand should be used. When replacing a crystal diode, proper polarity must be observed.

Schematic circuit diagrams, with critical voltage check points indicated, and complete replacement parts lists are given to facilitate servicing. If realignment becomes necessary (e.g., after tube replacement), follow the step-by-step procedure.

TEST EQUIPMENT REQUIRED

1 Laboratory Type Oscilloscope

1 Marker Generator (Jerrold Models CM-6, CM-10, or equivalent)

1 Field Strength Meter (capable of covering the oscillator frequency range)

1 Variable Attenuator (Jerrold Models AV-50, AV-50-2, or equivalent)

1 R. F. Voltmeter or bolometer

1 Vacuum Tube Voltmeter

ALIGNMENT OF OSCILLATOR HEADS H-71A, H-72A AND H-73A

A. SETUP

Block diagram Fig. 2 shows how to setup Model 707-D and test equipment after a tube replacement in the oscillator plug-in unit.

B. STEP-BY-STEP PROCEDURE

1. Switch on all the instruments; allow 20 to 30 minutes for proper warm-up.

2. Set field strength meter to 54 mc.

3. Set Function Switch on field strength meter to SIGNAL STRENGTH.

4. Set all attenuator switches to ON.

5. Set SW WIDTH control on 707-D to maximum range.

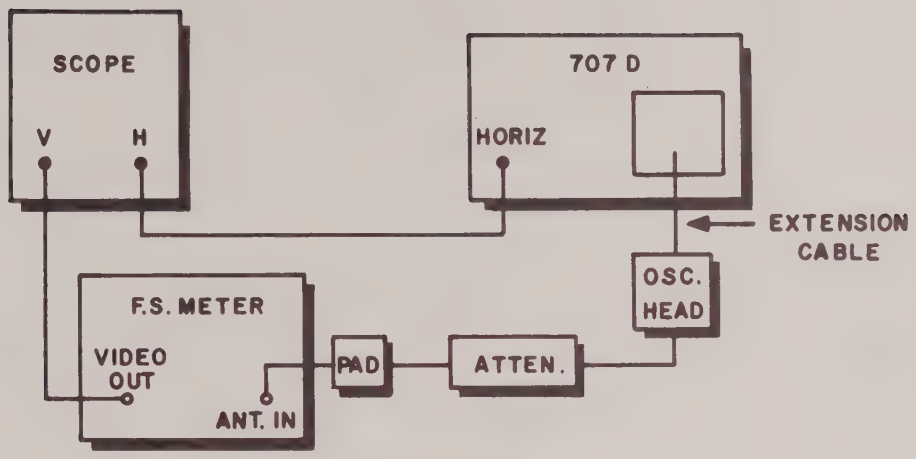


Figure 2. Setup for Alignment of Oscillator

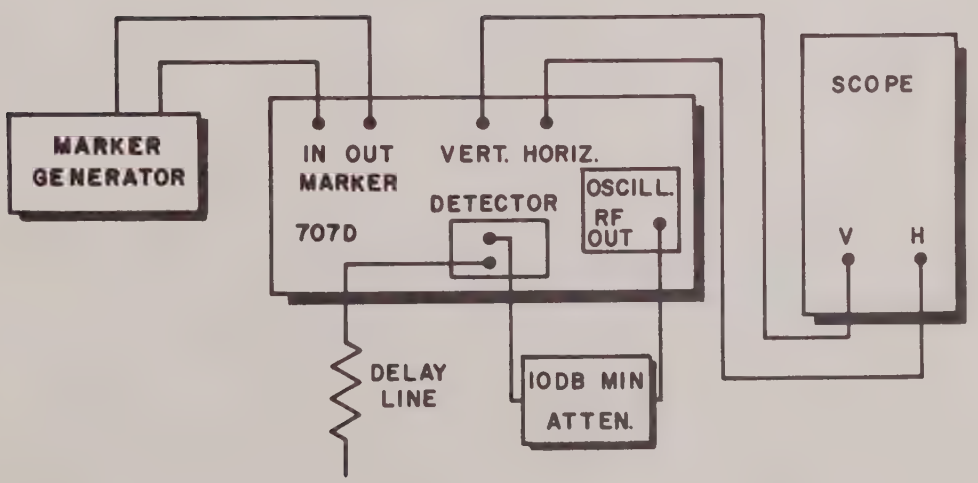


Figure 3. Setup to Align for Linearity

6. Set ALC on 707-D to 3 v output (to 1 v for Model H-72A).

7. Set CENTER FREQ control on 707-D to ON position.

8. Set BLANKING switch on 707-D to ON position.

9. Set AUTO-MAN switch on 707-D to AUTO position.

10. Set SWEEP SPEED FINE control on 707-D to max. range.

11. Set SW/MIN-SW/SEC switch on 707-D to .6-6 range (0.6 to 6.0 sweeps per second).

12. Set SW WIDTH control on 707-D to present optimum width of spike trace on scope, while spike amplitude remains the same.

13. Set the frequency control on the field strength meter to 108 mc; observing the scope, the spike should be seen going off the screen.

14. Reduce attenuation until the spike reappears on the CRT screen.

15. Adjust the variables R-2, C36 and C37 on the H-71A head to obtain minimum amplitude of spike, first by setting C36 and C37 at minimum position, then (slowly turning one capacitor) observe whether the spike amplitude increases or decreases. If an increase is noted, reset the capacitor to minimum, then adjust the other capacitor for minimum spike amplitude. After that, adjust R-2 to further decrease the spike amplitude. Repeat this step until absolute minimum spike amplitude is obtained.

NOTE: The alignment procedures for Models H-72A and H-73A are the same as that for Model H-71A except that the field strength meter settings are to be made first at 100 mc, then at 200 mc for Model H-72A, and — for Model H-73A — a detector capable of registering down into the low frequency range should be used instead of the FSM.

If a Jerrold field strength meter Model 704B is used, the Model TK-704 100 mc adapter will permit the FSM to measure frequencies down to 8 mc. Set the FSM first at 125 mc, then for harmonic at 150 mc. The max. input signal voltage to the adapter should not exceed 3000 microvolts; where it does, use an appropriate pad or attenuator.

ALIGNMENT OF SWEEP GENERATOR

A. SWEEP CIRCUIT

NOTE: Only an oscilloscope is required.

1. Connect VERT input of scope to HORIZ output connector J6 at rear of 707-D chassis.
2. Set AUTO-MAN switch on 707-D to AUTO position.
3. Set PYR-SAW switch on 707-D to PYR position.
4. Set SW/MIN - SW/SEC switch on 707-D to 6-60 range.
5. Set scope vertical sensitivity to 40 volts full scale.
6. Adjust SWING potentiometer R6 on 707-D chassis to obtain 40-volt sweep trace on scope.
7. Adjust SYM control R27 on 707-D chassis to obtain perfect symmetry of trace.
8. Adjust DC REF control R9 on 707-D chassis to adjust level of pyramidal output.

B. ALIGNMENT FOR LINEARITY

NOTE: Model 707-D is factory-adjusted for best overall linearity for all three plug-in oscillator models. If it is desired to further optimize linearity for a particular oscillator head, follow the step-by-step procedure given here.

1. Set up Model 707-D and test equipment as shown in block-diagram Fig. 3.
2. Switch on all instruments and allow 20 - 30 minutes for warm-up.
3. Set AUTO-MAN switch on Model 707-D to AUTO position.
4. Set BLANKING on scope to ON position.
5. Set PYR-SAW switch on Model 707-D to PYR position.
6. Set ALC control on Model 707-D front panel to provide 3 volts r.f. output on meter.
7. Adjust potentiometers LIN A and LIN B on Model 707-D chassis to obtain maximum linearity of ripple pattern on scope.

C. ALIGNMENT OF METERING CIRCUIT

NOTE: Readjustment of this circuit becomes necessary whenever tube V9 is replaced. For readjustment it is preferable to use a standard H-71A or H-73A oscillator head which provide a 3-volt output; however, readings for an H-72A head are also given.

1. Plug oscillator head into the generator power supply.

2. Connect dc voltmeter with positive lead to terminal 7 and negative lead to terminal 8 on multi-pin connector J2 at generator chassis (rear of housing which shields the oscillator).

3. Energize the unit and allow about 15 minutes warm-up time. Do not energize before having first connected the voltmeter; terminals 7 and 8 on J2 are -150 volts above ground potential.

4. Set sweep mode to MAN; set blanking to FWD or REV, whichever provides the blanked output.

5. Adjust meter ZERO control on front panel to indicate zero on the meter.

6. Set BLANKING switch to OFF position.

7. Adjust ALC control on front panel to give a reading of 1.5 volts on the dc voltmeter (or 1.0 vdc for Model H-72A).

8. Adjust METER SET control on chassis to give a 3-volt reading on 4-volt scale of the front panel meter M1 (or 1.0 volt reading on 2-volt scale for Model H-72A). Lock METER SET control.

9. To adjust METER INPUT control on oscillator head, measure the r.f. output with an r.f. voltmeter. Adjust ALC control on front panel of 707-D for a 3-volt rms output (1-volt rms for Model H-72A) and adjust METER INPUT control (accessible from chassis bottom) for proper indication on front panel meter M1 (3 volts on 4-volt scale for H-71A and H-73A, 1 volt on 2-volt scale for H-72A).

REPLACEMENT PARTS LIST

SWEEP FREQUENCY GENERATOR — MODEL 707-D

(See Dwg. 861-538)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	CAPACITORS			
1	C1A, B	1	200+80 UF, 350V, dual electrolytic	127-609
2	C2	1	20 UF, 450V, electrolytic	127-025
3	C3, C4	2	40 UF, 450V, electrolytic	127-505
4	C5	1	100 PF + 10%, 500 V	123-105
5	C6, C9, C19	3	1 UF, 200 V	125-019
6	C7	1	0.01 UF, 400 V	125-013
7	C8	1	0.01 UF, 200 V	125-002
8	C10*	1	10 UF, bathtub	125-047
9	C13	1	40 UF, 250 V, electrolytic	127-004
10	C14	1	0.02 UF, 500 V, GMV, ceramic disc	124-034
11	C16	1	1000 PF, 500 V, GMV, ceramic disc	123-115
12	C17	1	150 PF + 10%, 500 V	123-107
13	C18	1	3000 PF, 600 V	125-028
			NOTE: C15 not assigned	
	CHOKE			
14	L1	1	2.3 H, 150 MA, laminated iron core choke	B143-105
	FUSE			
15	F1	1	2A/115V or 1A/230V, SLO-BLO	101-345 or 101-352
	METER			
16	M1	1	0-200 UA, DC	171-206
	PILOT LAMPS			
17	DS1, 2	2	6-8 V, 150 MA, #47	102-002

* In some units 3 capacitors, C10 (2 MF) C11 (4MF) and C12 (4 MF) are used for a composite capacitance of 10 MF.

REPLACEMENT PARTS LIST

SWEEP FREQUENCY GENERATOR — MODEL 707-D (Cont)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	PLUGS, RECEPTACLES			
18	P1	1	Line cord and 3-pin plug	659-116
19	Shipped loose	1	Grounding adapter	180-105
20	J1	1	6-pin socket	182-114
21	J2	1	16-pin socket	B183-209
22	J3, 6, 7, 8	4	BNC chassis fitting	188-103
23	P2, 3	2	Banana plug	184-020
	RECTIFIERS			
24	CR1, 2, 3	3	Silicon diode, CER72ST	137-711
	RESISTORS			
25	R1, 4	2	1 K, 10 W, wirewound	113-005
26	R2	1	1.25 K, 20 W, wirewound	113-013
27	R3, 81	2	220 Ohms, ± 10%, 1 W	112-279
28	R5	1	1.75 K, 10 W, wirewound	113-008
29	R6, 9, 27	3	500 K, 2 W, potentiometer	118-030
30	R7, 28, 72, 75	3	1 Meg. ± 5%, 1/2 W	112-737
31	R8	1	3.6 Meg. ± 5%, 1/2 W	112-809
32	R10, 12	2	68 K ± 5%, 1/2 W	112-590
33	R11, 19, 29, 79	4	3 Meg. ± 5%, 1/2 W	112-797
34	R13, 16	2	3.3 Meg. ± 5%, 1/2 W	112-800
35	R14, 32, 33, 34, 64	5	1 K ± 5%, 1/2 W	112-359
36	R15, 18	2	150 K ± 5%, 1/2 W	112-632
37	R17, 23, 39, 46, 47	5	10 K ± 5%, 1/2 W	112-485
38	R20, 42	2	200 K ± 5%, 1/2 W	112-650
39	R21, 43, 48	3	100 K ± 5%, 1/2 W	112-611
40	R22	1	91 K ± 5%, 1/2 W	112-608
41	R24	1	12 K ± 5%, 1/2 W	112-497
42	R25	1	250 K, 2 W, potentiometer	118-120
43	R26	1	22 K, ± 5%, 1/2 W	112-527
44	R30, 36	2	510 K, ± 5%, 1/2 W	112-704
45	R31, 41	2	5 Meg. 2 W, potentiometer	118-054
46	R35, 38	2	500 Ohms, ± 10%, 2 W, potentiometer	118-053
47	R37, 74	2	2.2 Meg. ± 5%, 1/2 W	112-779
48	R40	1	500 K, ± 10%, 2 W, potentiometer	118-056
49	R44	1	1 Meg. 2 W, potentiometer	118-055
50	R45	1	100 K, 2 W, potentiometer	118-062
51	R49	1	51 K ± 5%, 1/2 W	112-578
52	R50, 59, 60, 66, 73	5	200 Ohms, ± 5%, 1/2 W	112-272
53	R51	1	470 Ohms, ± 5%, 1/2 W	112-317
54	R52A, B	1	10 K, 100 K, 2 W dual potentiometer	118-115
55	R53, 70	2	33 K ± 5%, 1/2 W	112-548
56	R54	1	3.3 K ± 5%, 1/2 W	112-422

REPLACEMENT PARTS LIST

SWEEP FREQUENCY GENERATOR — MODEL 707-D (Cont)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	RESISTORS (Cont)			
57	R55	1	10 K, 2 W, potentiometer	118-026
58	R56	1	500 Ohms ± 10%, 2 W, potentiometer	118-103
59	R57	1	22 K ± 10%, 2 W	112-532
60	R58	1	2 Meg. ± 5%, 1/2 W	112-776
61	R61	1	2.7 Meg. ± 5%, 1/2 W	112-791
62	R62	1	56 K ± 5%, 1/2 W	112-581
63	R63	1	120 K ± 5%, 1/2 W	112-623
64	R65	1	1.8 K ± 5%, 1/2 W	112-392
65	R67	1	82 K ± 5%, 1/2 W	112-602
66	R68	1	100 Ohms ± 5%, 1/2 W	112-233
67	R69	1	220 K ± 5%, 1/2 W	112-653
68	R71	1	47 K ± 5%, 1/2 W	112-569
69	R76	1	180 K ± 5%, 1/2 W	112-644
70	R77, 78	2	270 Ohms ± 5%, 1/2 W	112-287
71	R80	1	47 Ohms ± 5%, 1/2 W	112-191
72	R82, 83, 85, 86	4	22 Ohms ± 10%, 1/2 W	112-152
73	R84	1	6.8 Meg. ± 5%, 1/2 W	112-842
74	R87	1	330 K ± 5%, 1/2 W	112-674
75	R88	1	38.3 K ± 1%, 1 W	115-117
76	R89	1	1 K ± 1%, 1/2 W	115-128
77	R90	1	15 K ± 10%, 2 W	112-511
78	R91	1	1.1 K ± 5%, 1/2 W	112-368
	SWITCHES			
79	S1, 3	2	SPST, toggle	162-001
80	S2	1	SPDT, 3-position, toggle	162-013
81	S4	1	DPDT, toggle	162-008
82	S5	1	4-position, single-deck, wafer	B161-116
83	S6	1	SPDT, toggle	162-007
	TRANSFORMER			
84	T1	1	Line transformer, 115/230 V	C141-188
	TUBES			
85	V1, 2, 3	3	OD3/VR150 voltage regulator	132-101
86	V4, 5	2	12AT7, twin triode, 9-pin miniature	131-400
87	V6, 12	2	6U8, triode-pentode, 9-pin miniature	131-320
88	V7	1	6CL6, pentode, 9-pin miniature	131-327
89	V8	1	6K6, pentode, octal	131-328
90	V9	1	12AX7, twin triode, 9-pin miniature	131-404
91	V10, 11	2	6AU6, pentode, 7-pin miniature	131-308
92	V13	1	12BY7A, pentode, 9-pin miniature	131-403

REPLACEMENT PARTS LIST

PLUG-IN DETECTOR MODEL D-51

(See Dwg. 861-366)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
1	C-101	1	1000 MMF, mica button	129-153
2	C-102	1	47 MMF, feed-thru	129-202
3	C-103	1	1000 MMF, feed-thru	129-200
4	CR-101, -102	2	1N34, Germanium diodes	139-112
5	J-102, -103	2	Banana jacks	185-107
6	J-104, -105	2	UG-657/U type chassis fittings	188-103
7	R-101, -103	2	300 Ohms, + 5%, 1/2 W	112-293
8	R-102	1	100 K, $\pm$ 10%, 1/2 W	112-614

REPLACEMENT PARTS LIST

PLUG-IN OSCILLATOR MODEL H-71A

(See Dwg. 861-277)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	CAPACITORS*			
1	C-2	1	1.1 MMF ± 0.1 MMF, 500 V, ceramic	121-071
2	C-3	1	3 MMF ± 0.1 MMF, 500 V, ceramic	121-072
3	C-4	1	4.7 MMF ± 0.1 MMF, 500 V, ceramic	121-073
4	C-5	1	6.2 MMF ± 0.1 MMF, 500 V, ceramic	121-074
5	C-6	1	9.1 MMF ± 0.1 MMF, 500 V, ceramic	121-076
6	C-7	1	12 MMF ± 2%, 500 V, ceramic	121-012
7	C-8, -16, -17	3	15 MMF ± 2%, 500 V, ceramic	121-013
8	C-9	1	20 MMF ± 2%, 500 V, ceramic	121-014
9	C-10	1	24 MMF ± 2%, 500 V, ceramic	121-016
10	C11, -13, 14, 15, 20, 21, 30, -32, -33, -34, -35	11	1000 MMF, 500 V, ceramic, feed-thru	129-200
11	C-12, 31	2	47 MMF, 500 V, ceramic, feed-thru	129-202
12	C-22, -24, -26	3	8.2 MMF ± 0.25 MMF, 500 V, ceramic	121-075
13	C-23	1	27 MMF ± 2%, 500 V, ceramic	121-018
14	C-25	1	1.2 MMF ± 0.25 MMF, 500 V, ceramic	121-038
15	C-27	1	13 MMF ± 2%, 500 V, ceramic	121-043
16	C-28	1	0.6 - 10 MMF, H. F. Trimmer, Model HFT-65A	B821-203
17	C-29	1	1000 MMF, 500 V, ceramic, standoff	129-199
18	C36, 37	2	0.7 - 3.0 MMF, 350 V, trimmer	128-505
	COILS			
19	L5	1	12 MICRO-H, choke coil	157-018
20	L-12	1	Saturating coil	157-026
	CONNECTORS			
21	J-1	1	BNC chassis fitting	188-103
22	J-2	1	16-pin plug	B183-210
	DIODES			
23	CR1, 2	2	1N34, Germanium	139-112
	NEON BULBS			
24	DS-1, -2	2	NE-2, neon glow	102-003

* C18, C19, not assigned.

REPLACEMENT PARTS LIST

PLUG-IN OSCILLATOR MODEL H-71A (Cont)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	RESISTORS			
25	R-1, -3	2	2 K ± 5%, 1/2 W	112-398
26	R-2	1	100 Ohms ± 10%, 2 W, potentiometer	118-025
27	R-4	1	93.1 Ohms ± 1%, 1/2 W	115-151
28	R5	1	2.2 K ± 10%, 1/2 W	112-404
29	R-6, -9, -10	3	1 K ± 5%, 1/2 W	112-359
30	R-7	1	49.9 Ohms ± 1%, 1/2 W	115-137
31	R-8	1	270 Ohms ± 5%, 1/2 W	112-287
32	R11	1	100 K ± 10%, 2 W, potentiometer	118-022
	SWITCH			
33	S1	1	11-position, rotary, wafer	161-125
	TRANSFORMER			
34	T-1	1	Oscillator transformer	B141-148
	TUBES			
35	V-1, -2	2	12BY7A, pentode, 9-pin miniature	131-403

REPLACEMENT PARTS LIST

PLUG-IN OSCILLATOR MODEL H-72A

(See Dwg. 861-368)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	CAPACITORS*			
1	C201	1	1.8 MMF ± 0.25 MMF, 500 V, ceramic	121-039
2	C202	1	2.7 MMF ± 0.25 MMF, 500 V, ceramic	121-040
3	C203	1	4.7 MMF ± 0.1 MMF, 500 V, ceramic	121-073
4	C204, 222	2	7.5 MMF ± 0.1 MMF, 500 V, ceramic	121-041
5	C205	1	10 MMF ± 2%, 500 V, ceramic	121-011
6	C206, 221	2	15 MMF ± 2%, 500 V, ceramic	121-013
7	C207	1	18 MMF ± 2%, 500 V, ceramic	121-046
8	C208	1	22 MMF ± 2%, 500 V, ceramic	121-015
9	C209	1	33 MMF ± 2%, 500 V, ceramic	121-020
10	C210	1	50 MMF ± 2%, 500 V, ceramic	121-025
11	C211, 228	2	47 MMF, 500 V, ceramic, feed-thru	129-202
12	C212, 213, 215, 218, 227, 230, 231	7	1000 MMF, 500 V, ceramic, feed-thru	129-200
13	C214, 217	2	0.7 -3 MMF, 350 V, trimmer	128-505
14	C216	1	330 MMF ± 10%, 600 V	123-110
15	C219, 220	2	5 MMF ± 0.25 MMF, 500 V, ceramic	121-008
16	C223	1	5.1 MMF ± 0.1 MMF, 500 V, ceramic	121-051
17	C224	1	0.6 - 10 MMF, h.f. trimmer Model HFT-65A	B821-203
18	C225	1	1000 MMF, 500 V, ceramic, standoff	129-199
19	C229	1	0.02 MF ± 10%, 350 V	124-034
	COILS			
20	L201	1	Saturating coil	157-026
21	L202	1	Choke coil	157-006
	CONNECTORS			
22	J201	1	BNC chassis fitting	188-103
23	J202	1	16-pin plug	183-210
	DIODES			
24	CR201, 202	2	1N34, Germanium	139-112
	NEON BULBS			
25	DS201, 202	2	NE-2, neon glow	102-003

* C226 not assigned.

REPLACEMENT PARTS LIST

PLUG-IN OSCILLATOR MODEL H-72A (Cont)

ITEM	CIRCUIT DESCRIPTION	QTY.	DESCRIPTION	JERROLD PART NO.
	RESISTORS*			
26	R201, 203	2	4.7 K ± 5%, 1/2 W	112-443
27	R202	1	100 Ohms ± 10%, 2 W, potentiometer	118-025
28	R204	1	75 Ohms ± 1%, 1/2 W	115-108
29	R205	1	2.2 K ± 10%, 1/2 W	112-404
30	R206	1	180 Ohms ± 5%, 1/2 W	112-266
31	R207	1	120 Ohms ± 5%, 1/2 W	112-245
32	R208	1	49.9 Ohms ± 1%, 1/2 W	115-137
33	R209, 210	2	1 K ± 5%, 1/2 W	112-359
34	R212	1	100 K ± 10%, 2 W, potentiometer	118-022
35	R213	1	100 K ± 10%, 1/2 W	112-614
	SWITCH			
36	S201	1	11-position, rotary, wafer	161-125
	TRANSFORMER			
37	T201	1	Oscillator transformer	144-033
	TUBE			
38	V201A, B	1	6922, twin triode	131-602

* R211 not assigned.

REPLACEMENT PARTS LIST

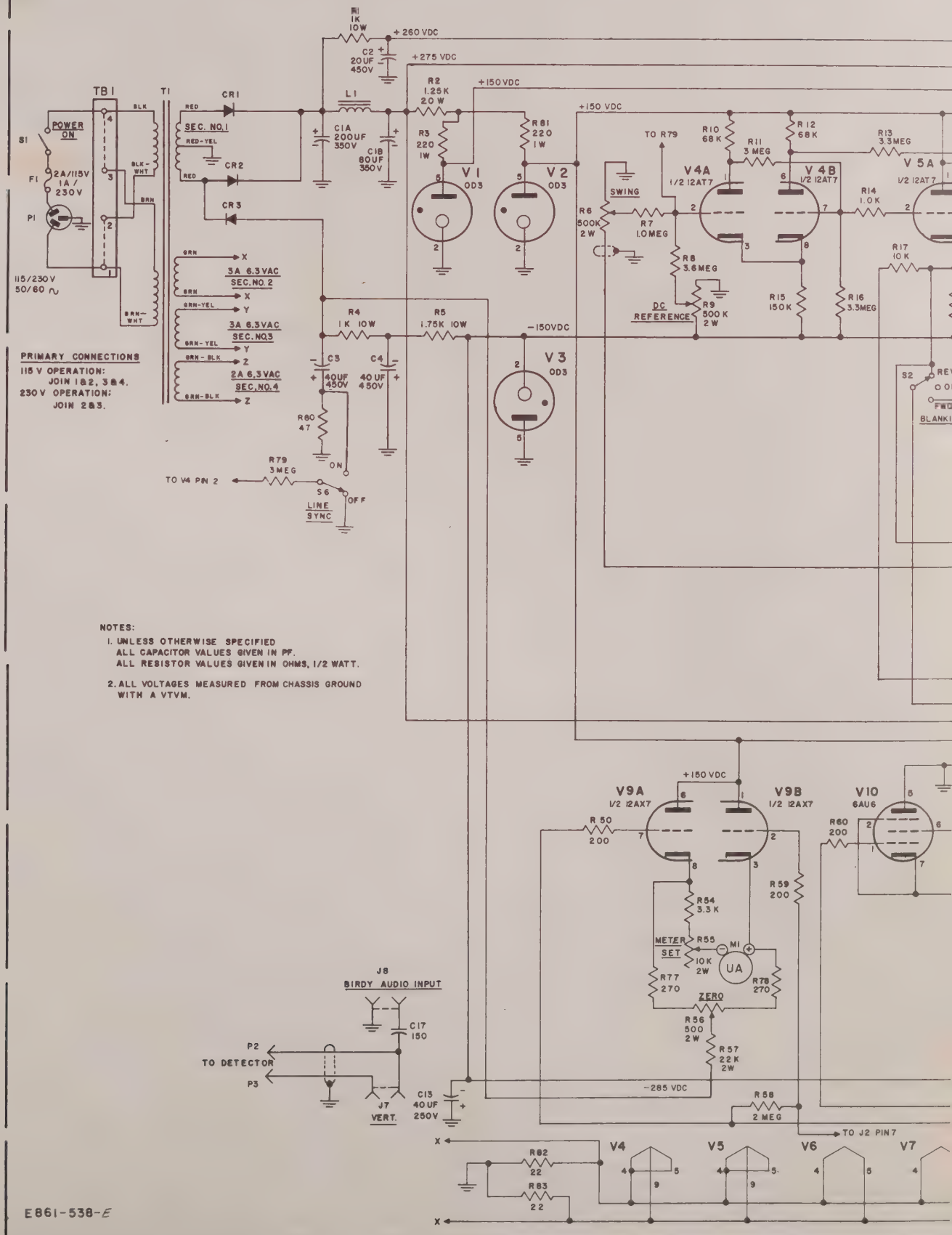
PLUG-IN OSCILLATOR MODEL H-73A

(See Dwg. 861-367)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	CAPACITORS			
1	C301	1	0.5 MMF ± 0.25 MMF, 600 V	121-001
2	C302	1	1.0 MMF ± 0.25 MMF, 600 V	121-003
3	C303	1	2.7 MMF ± 0.25 MMF, 500 V, ceramic	121-040
4	C304	1	4.7 MMF ± 0.1 MMF, 500 V, ceramic	121-007
5	C305	1	6.8 MMF ± 0.1 MMF, 500 V, ceramic	121-009
6	C306	1	12 MMF ± 2%, 500 V, ceramic	121-012
7	C307	1	25 MMF ± 2%, 500 V, ceramic	121-017
8	C308	1	36 MMF ± 2%, 500 V, ceramic	121-021
9	C309	1	62 MMF ± 2%, 500 V, ceramic	121-028
10	C310	1	200 MMF ± 5%, 500 V, mica	126-007
11	C311, 331	2	47 MMF, 500 V, ceramic, feed-thru	129-202
12	C312, 314, 315, 316, 321, 322, 332, 333, 334, 335, 336	11	1000 MMF, 500 V, ceramic, feed-thru	129-200
13	C313	1	0.01 MF ± 10%, 350 V	124-031
14	C317, 320	2	0.7-3 MMF, 350 V, trimmer	128-505
15	C318, 319	2	1000 MMF ± 20%, 600 V, ceramic	123-115
16	C323, 324, 327	3	15 MMF ± 2%, 500 V, ceramic	121-013
17	C325	1	47 MMF ± 2%, 500 V, ceramic	121-024
18	C326	1	68 MMF ± 2%, 500 V, ceramic	121-029
19	C328	1	1.2 MMF ± 0.5 MMF, 500 V, ceramic	121-038
20	C329	1	0.6-10 MMF, H. F. trimmer Model HFT-65A	B821-203
21	C330	1	1000 MMF, 500 V, ceramic, standoff	129-199
	COILS			
22	L301	1	Saturating coil	157-026
23	L302	1	300 MICRO-H, choke coil	157-030
	CONNECTORS			
24	J301	1	BNC chassis fitting	188-103
25	J302	1	16-pin plug	B183-210
	DIODES			
26	CR301, 302	2	1N34, Germanium	139-112
	NEON BULBS			
27	DS301, 302	2	NE-2, neon glow	102-003

PLUG-IN OSCILLATOR MODEL H-73A (Cont)

ITEM	CIRCUIT DESIGNATION	QTY.	DESCRIPTION	JERROLD PART NO.
	RESISTORS			
28	R301, 303	2	3 K ± 5%, 1/2 W	112-419
29	R302	1	100 Ohms ± 10%, 2 W, potentiometer	118-025
30	R304	1	98.5 Ohms ± 1%, 1/2 W	115-128
31	R305	1	2.2 K ± 10%, 1/2 W	112-404
32	R306	1	2 K ± 5%, 1/2 W	112-398
33	R307	1	680 Ohms ± 5%, 1/2 W	112-338
34	R308	1	49.9 Ohms ± 1%, 1/2 W	115-137
35	R309, 310	2	1 K ± 5%, 1/2 W	112-359
36	R311	1	50 K ± 10%, 2 W, potentiometer	118-024
37	R312	1	51 K ± 5%, 1/2 W	112-578
	SWITCH			
38	S301	1	11-position, rotary, wafer	161-125
	TRANSFORMER			
39	T301	1	Oscillator transformer	144-408
	TUBE			
40	V301, 302	2	12BY7A, pentode, 9-pin miniature	131-403

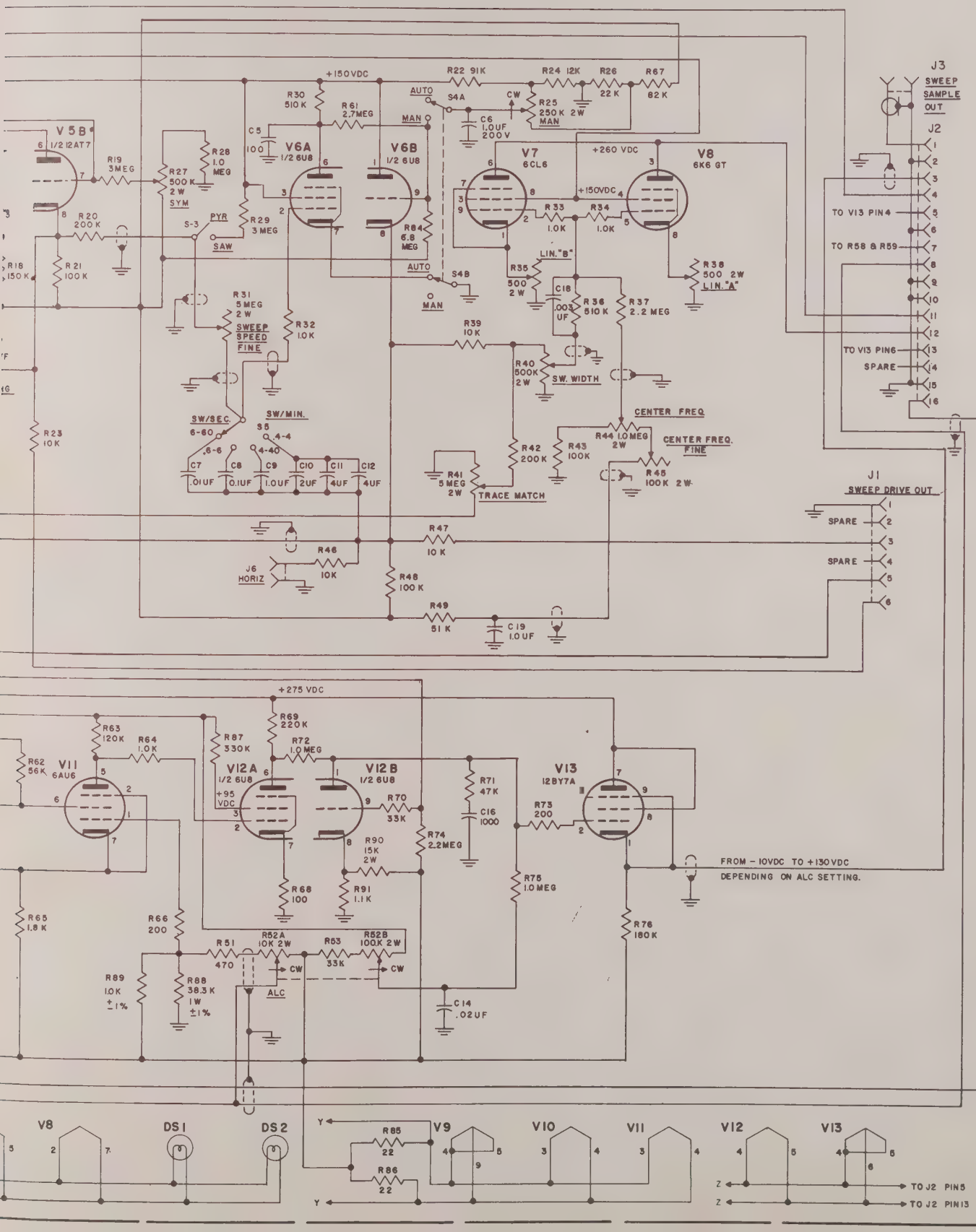

R1 1K 10W
+260 VDC
C2 20UF 450V
+275 VDC
+150VDC
R2 1.25K 20 W
R3 220 1W
R81 220 1W
+150 VDC
TB1
T1
CR1
CR2
CR3
L1
POWER ON
S1
F1 2A/115V 1A/230V
P1
115/230V 50/60
PRIMARY CONNECTIONS
115 V OPERATION: JOIN 1&2, 3&4.
230 V OPERATION: JOIN 2&3.
SEC. NO.1
C1A 200UF 350V
C1B 80UF 350V
3A 6.3VAC SEC. NO.2
3A 6.3VAC SEC. NO.3
2A 6.3VAC SEC. NO.4
R4 1K 10W
R5 1.75K 10W
-150VDC
C3 40UF 450V
C4 40 UF 450V
R80 47
R79 3MEG
TO V4 PIN 2
S6
ON
OFF
LINE SYNC
V1 0D3
V2 0D3
V3 0D3
TO R79
R10 68K
R11 3 MEG
R12 68K
R13 3.3MEG
V4A 1/2 12AT7
V4B 1/2 12AT7
V5A 1/2 12AT7
SWING
R6 500K 2W
R7 1.0MEG
R8 3.6MEG
R9 500 K 2W
DC REFERENCE
R14 1.0K
R15 150K
R16 3.3MEG
R17 10K
S2
NOTES:
1. UNLESS OTHERWISE SPECIFIED ALL CAPACITOR VALUES GIVEN IN PF. ALL RESISTOR VALUES GIVEN IN OHMS, 1/2 WATT.
2. ALL VOLTAGES MEASURED FROM CHASSIS GROUND WITH A VTVM.
+150 VDC
V9A 1/2 12AX7
V9B 1/2 12AX7
V10 6AU6
R50 200
R59 200
R60 200
R54 3.3 K
METER SET R55 10K 2W
MI
UA
R77 270
R78 270
ZERO
R56 500 2W
R57 22K 2W
J8 BIRDY AUDIO INPUT
C17 150
P2
TO DETECTOR
P3
J7 VERT.
C13 40 UF 250V
-285 VDC
R58 2 MEG
TO J2 PIN 7
V4
V5
V6
V7
R82 22
R83 22

E861-538-E

CHEMATIC

'EEP FREQUENCY GENERATOR

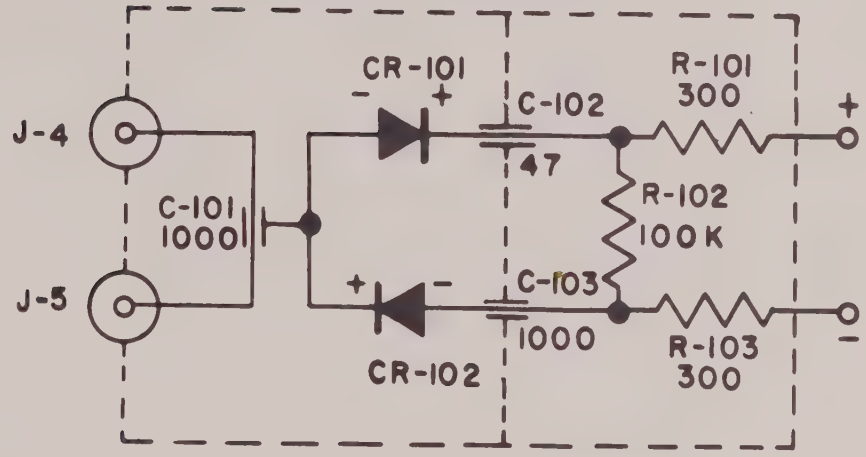
MODEL 707D



SCHEMATIC

SERIES-

PLUG-IN DETECTOR D-51



NOTES

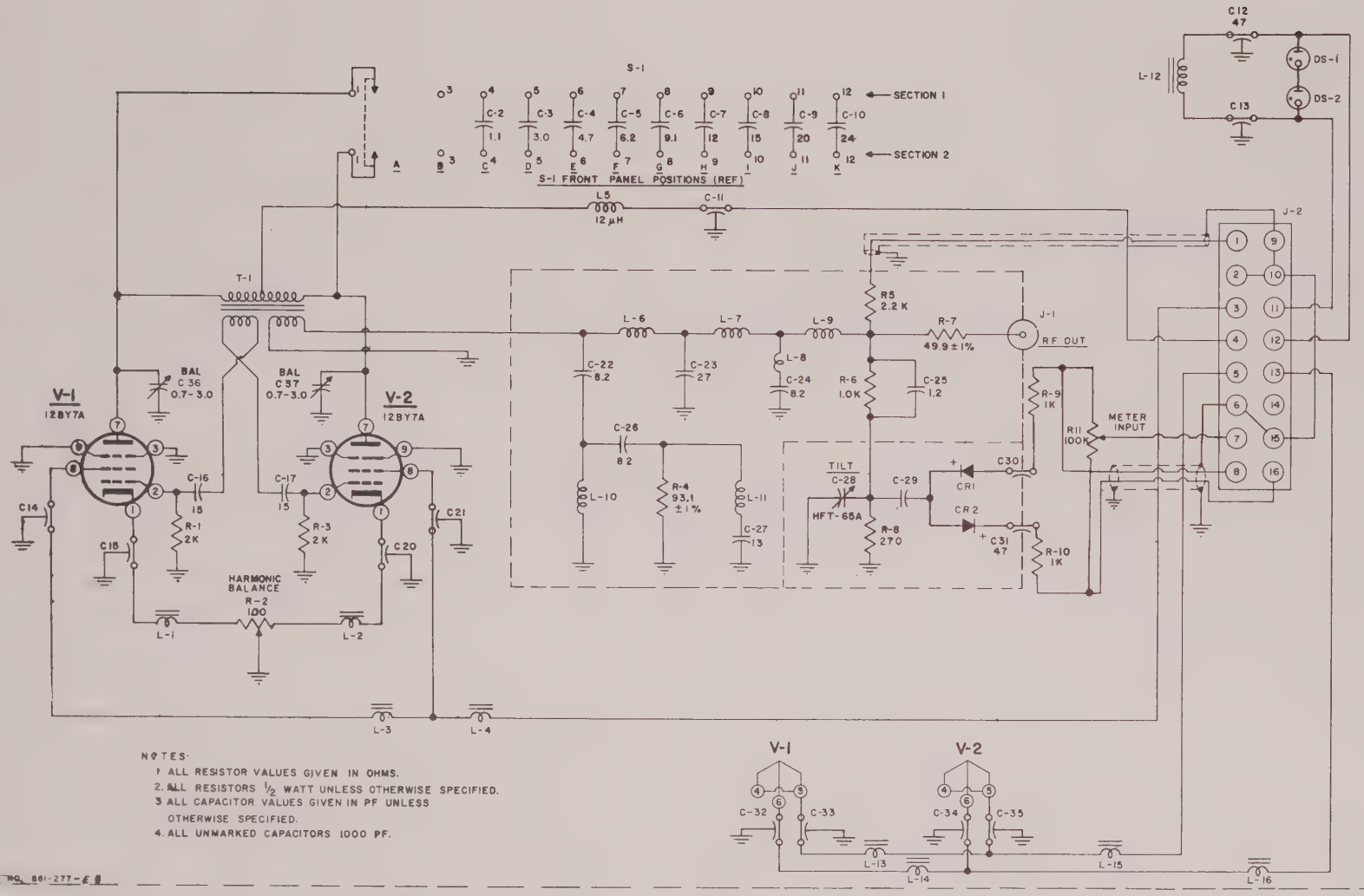
1. ALL RESISTOR VALUES GIVEN IN OHMS.
2. ALL RESISTORS 1/2 WATT UNLESS OTHERWISE SPECIFIED.
3. ALL CAPACITOR VALUES GIVEN IN MMF UNLESS OTHERWISE SPECIFIED.

DWG. NO. 861-366-0

SCHEMATIC

OSCILLATOR MODEL H-71A

SERIES -2



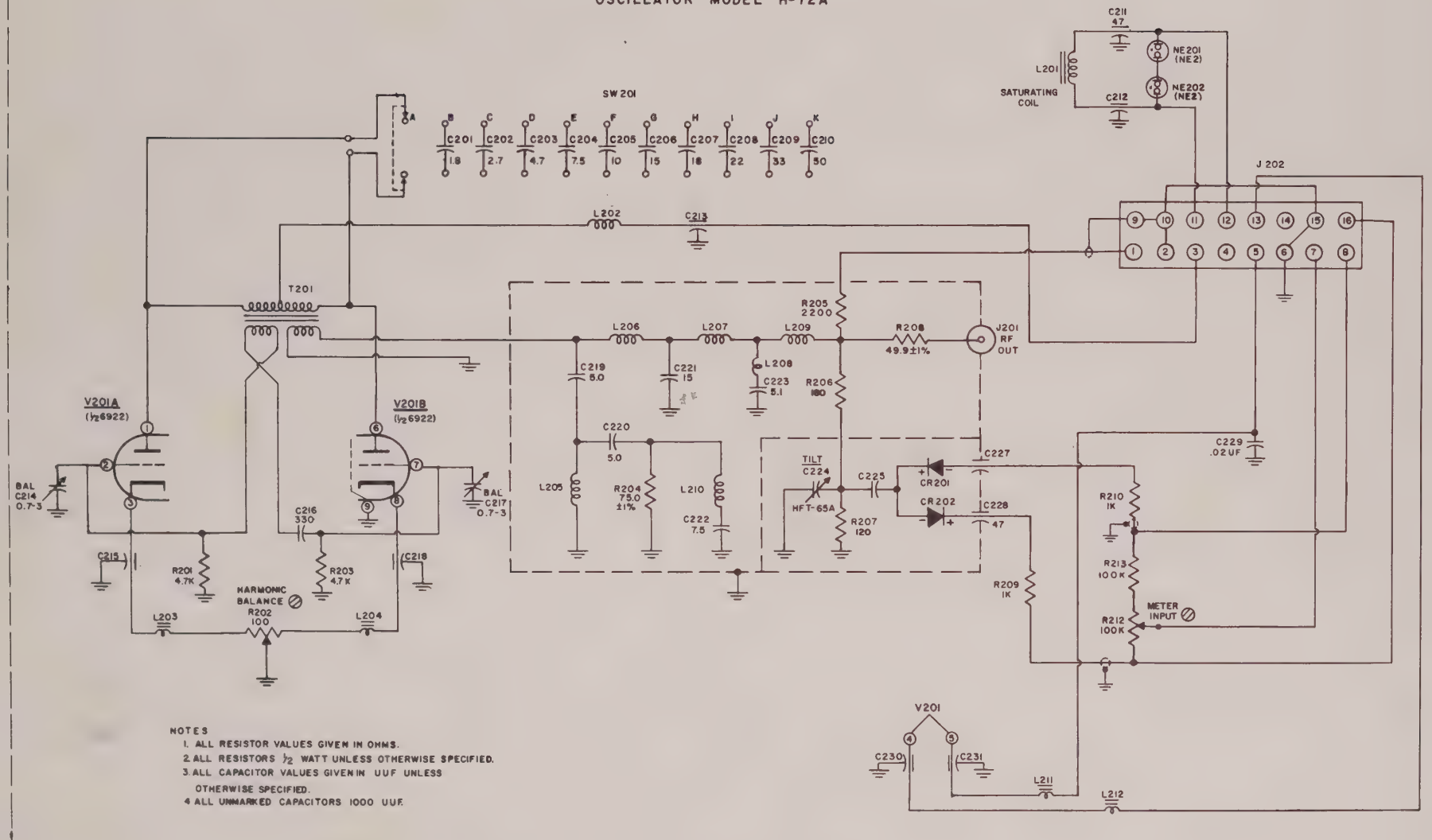
NOTES:

1. ALL RESISTOR VALUES GIVEN IN OHMS.
2. ALL RESISTORS ½ WATT UNLESS OTHERWISE SPECIFIED.
3. ALL CAPACITOR VALUES GIVEN IN PF UNLESS OTHERWISE SPECIFIED.
4. ALL UNMARKED CAPACITORS 1000 PF.

Dwg. No. 861-277-E B

SCHEMATIC

OSCILLATOR MODEL H-72A



NOTES

1. ALL RESISTOR VALUES GIVEN IN OHMS.
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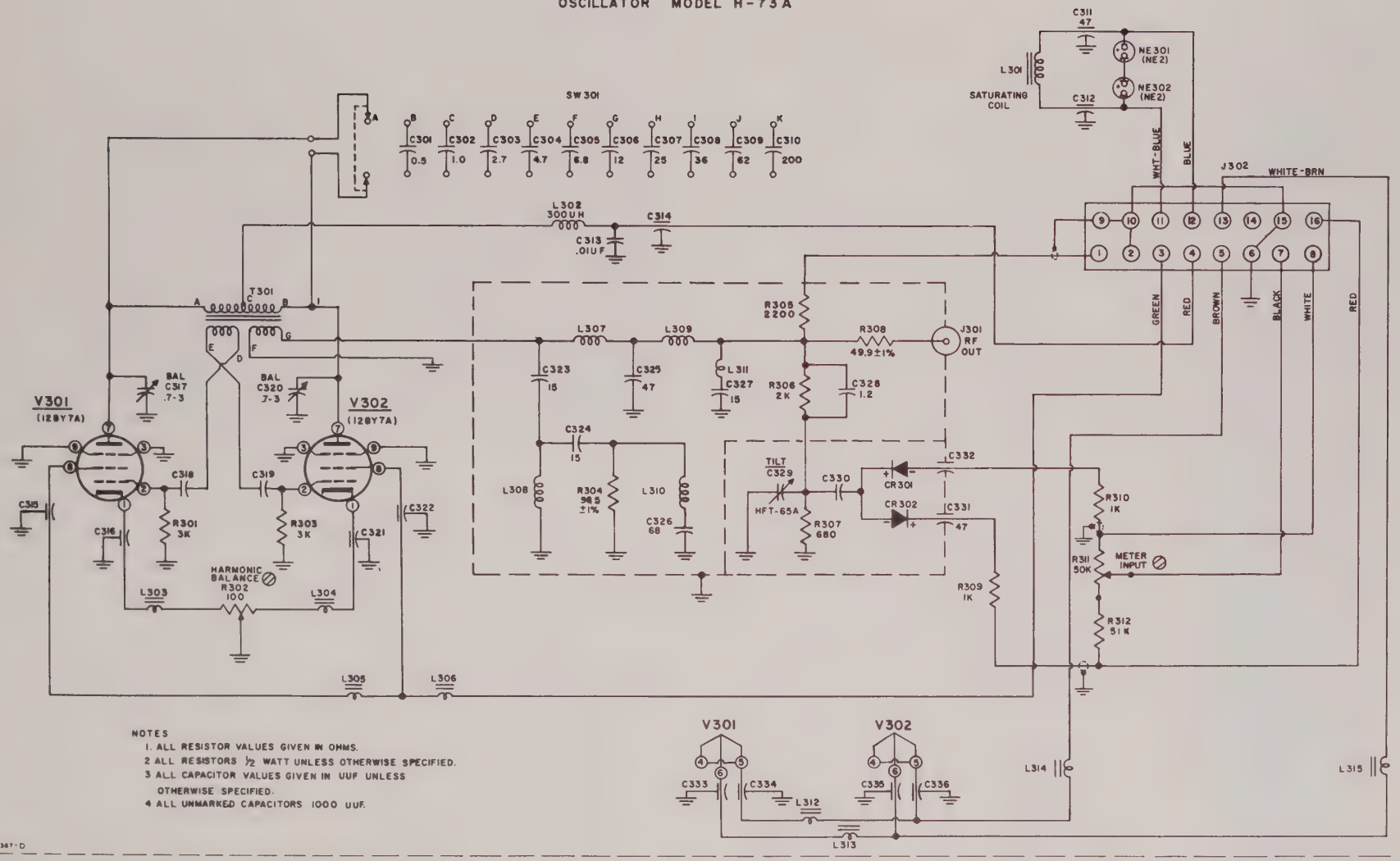
3. ALL CAPACITOR VALUES GIVEN IN UUF UNLESS OTHERWISE SPECIFIED.
4. ALL UNMARKED CAPACITORS 1000 UUF.

SCHEMATIC

OSCILLATOR MODEL H-73A



NOTES

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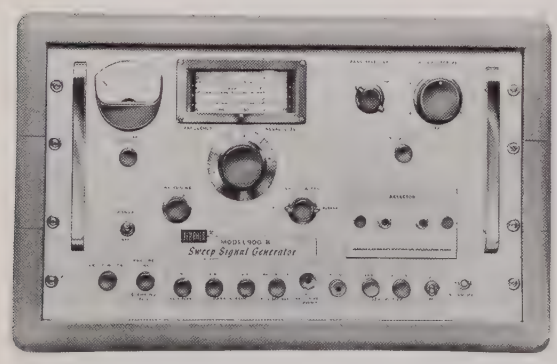
Wide Band Impedance Measurement

With a Sweep-Frequency Generator and Delay Line

by KEN SIMMONS

Vice President, Research and Development
Jerrold Electronics Corporation

The Sweep-Frequency, Delay Line method of testing was first used in the alignment and test of TV Broadcasting Antennas. Since then it has been widely used in the TV industry where impedance match over a wide band of frequencies was important. This article discusses the basic principles involved, and describes some techniques that allow an increase in measurement accuracy.



MODEL 900B
Sweep Frequency Generator

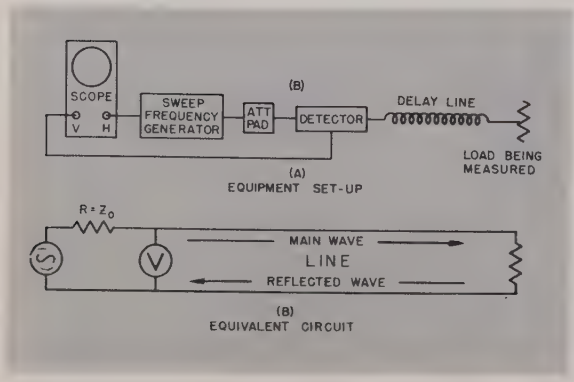


FIGURE 1

Equipment Set-Up

The basic equipment test set-up for wide-band impedance measurements is shown in Fig. 1 (A). An attenuator pad is generally connected between the sweep and the detector to isolate the generator from the varying impedance, and to ensure that the detector is fed from a well-matched source. The detector is needed when the sweep covers a frequency range above the upper response limit of the scope. The high-frequency voltage is rectified, and the detected output, a slowly-varying voltage representing the envelope of the high-frequency input, is applied to the oscilloscope.

The equivalent circuit of the test set-up is shown in Fig. 1 (B). A voltmeter (the detector and scope) measures the voltage at the junction between a matched source and a transmission line with a load on its far end. The voltage at this junction may be considered as being due to the sum of two waves: the *main wave* energy coming out of the sweep and going down the line, and the *reflected wave;* energy which has travelled down the line, been reflected from the load, and come back up the line.

The Delay Line

Terminated Line

With a well-designed sweep, the main-wave is made approximately constant with frequency. When the delay line is terminated in a resistance equal to its characteristic impedance, there is no reflected wave, so the scope shows a constant voltage. Fig. 2 illustrates a typical pattern. On the forward trace the output stays constant as the frequency changes across the sweep band. The reverse trace is "blanked"; the sweep output being keyed to zero to provide a reference line showing where zero output is.

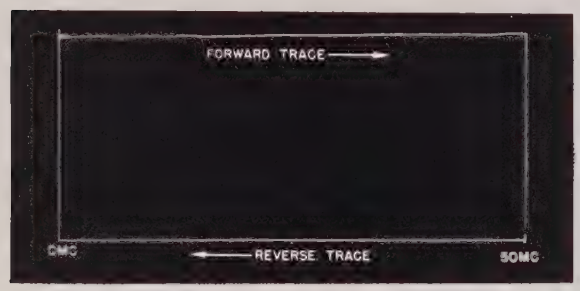


FIGURE 2

Open Circuit Line

When nothing is connected to the end of the line, this "open circuit" cannot dissipate, so all the energy striking it is reflected. When the line loss is small, the reflected wave at the detector has essentially the same amplitude as the main wave, and the total voltage there depends only on their phase relation. When they are in phase, they add to produce a voltage twice the "main wave only" condition. When they are 180° out of phase, they cancel to produce zero output. Fig. 3 illustrates the pattern with a generator sweeping 0 to 50 mc, and an open-circuited delay line ½ wavelength long at 50 mc. The reflected wave *at the open circuit* is in phase, at all frequencies, with the main wave. Near zero frequency, the effective length of the line is zero so the two waves are in phase and add at the detector, producing a maximum. At 25 mc, the line is ¼ wave-length long; the main wave shifts 90° in phase as it travels down the line to the end; and the reflected wave shifts another 90° on its way back, so the two components are 180° out of phase at the detector, giving a minimum at that frequency. At 50 mc, the reflected wave has travelled a full wavelength (½ wave down, ½ wave back) by the time it gets back to the detector, so the components add to a maximum.

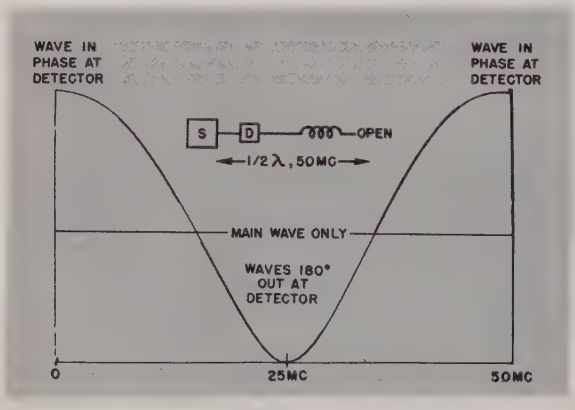


FIGURE 3

Short Circuit Line

Fig. 4 illustrates the pattern obtained by connecting a short circuit across the end of the line. This reverses the phase of the reflected wave so that the two components are 180° out at the short, and produce zero voltage there. At the detector, this produces a minimum at zero frequency (effective line length 0), a maximum at 25 mc., (reflected wave shifted 180° having travelled ½ wavelength extra), and a minimum again at 50 mc.

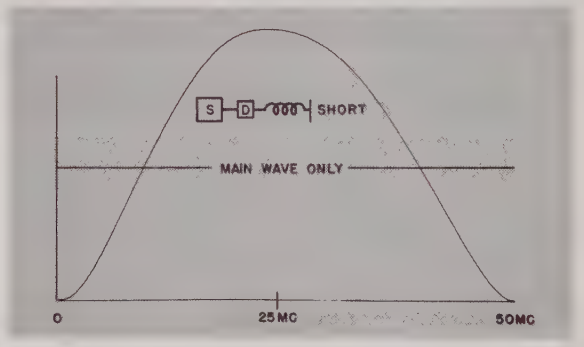


FIGURE 4

Longer Delay Lines

Fig. 5 shows the effect of doubling the length of the line, keeping the sweep-width the same. With an open-circuit (A) there are maxima at 0 frequency (0 line length), 25 mc (½ wavelength) and 50 mc (1 wavelength) and minima when the line length is ¼ and ¾ wavelength. A short-circuit termination gives the reverse pattern (B).

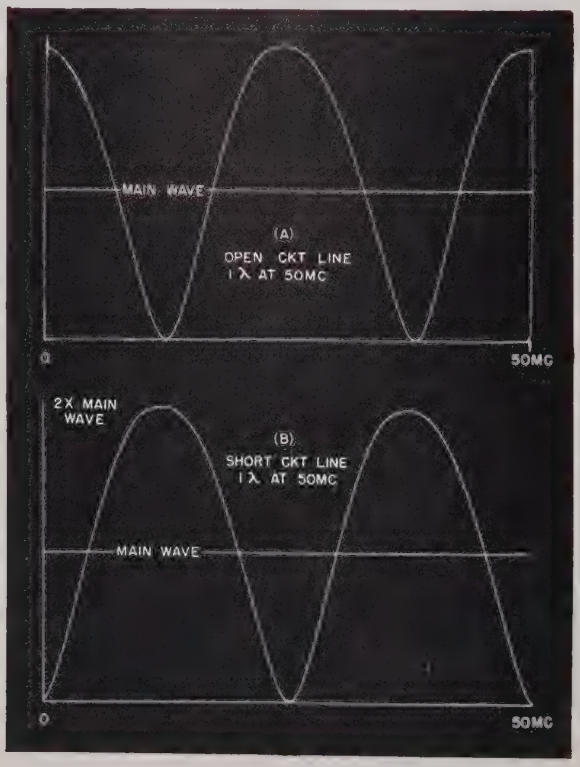


FIGURE 5

It is apparent that increasing the length of the delay line increases the number of ripples occuring in a given sweep-width. Specifically, the frequency change from one maximum (or minimum) to the next is equal to the frequency at which the line is one-half wavelength long. For a spacing of f mc between peaks, the line length in feet is $\frac{492\ d}{f}$ where d is the delay factor, the ratio of the speed with which a wave travels in the particular type cable, to the speed in air. d equals 0.67 for solid polyethylene insulated cables, and approximately 0.8 for polyfoam or other dielectrics with a high proportion of air insulation.

Since more ripples depict more clearly what happens in a given frequency band, in this respect, longer lines are preferred over shorter ones. One factor that limits the length that may be effectively used is the line attenuation. As the length is increased, more and more energy is lost in the line, so the reflected signal, as it shows up back at the detector, gets increasingly weaker. Fig. 6 (a) shows the ripple pattern resulting with a short-circuit termination, a sweep from 50 to 100 mc, and a high-grade delay line for 5 mc between peaks.

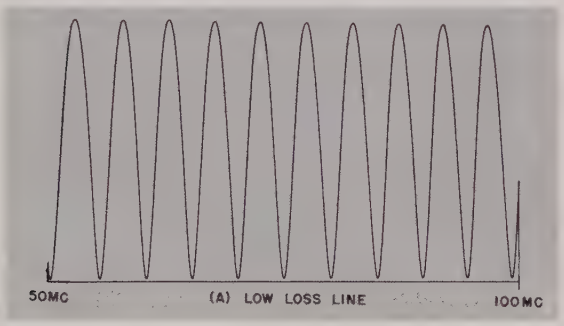


FIGURE 6A

The delay factor for this cable was 0.8, so its length, by the formula, would be $\frac{492 \times 0.8}{5}$ or 79 feet. Due to the greater length of this line, and the higher frequencies involved there is appreciable loss in the line. Although the reflected wave at the shorted end is equal to the main wave (reflection—100%), the main wave is stronger at the detector than it is at the far end, due to attenuation as it travels down the line, and the reflected wave is weaker at the detector than it is at the load, due to attenuation as it travels up the line from the load. This shows up on the ripple pattern in the fact that the minima do not quite go to zero, since the weakened reflected wave does not quite cancel the main wave.

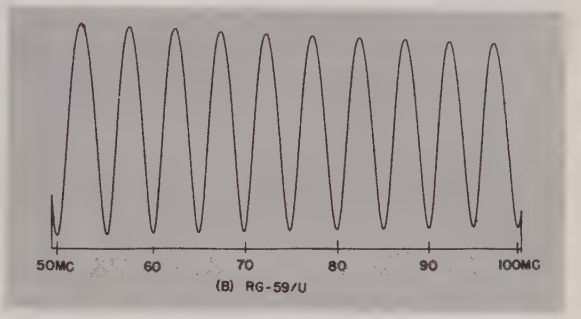
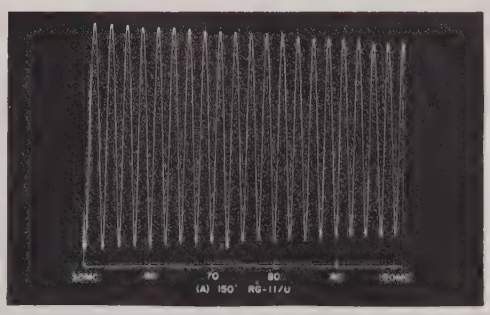


FIGURE 6B

Fig. 6 (b) shows the effect of using a cheaper cable (RG59/U) with higher loss. This line was cut for the same electrical length, but because its delay factor is 0.67, its physical length was shorter, 66 feet. Its greater attenuation shows up in the fact that the minima are still farther from zero. Notice that this effect is more pronounced at the high frequency end of the sweep, where the line loss is higher.

Fig. 7 (A) shows the ripple pattern from approximately 150' of RG11/U cable, shorted at the far end. Notice that the amplitude of the ripple pattern is about the same as that from the 66' piece of 59/U in Fig. 6 (B). The attenuation of RG11/U is about one-half that of RG59/U for a given length; so doubling the length just about cancels the attenuation improvement due to the larger cable. Fig. 7 (B) shows the pattern from about 500' of RG11/U showing the great reduction in ripple height at this length.



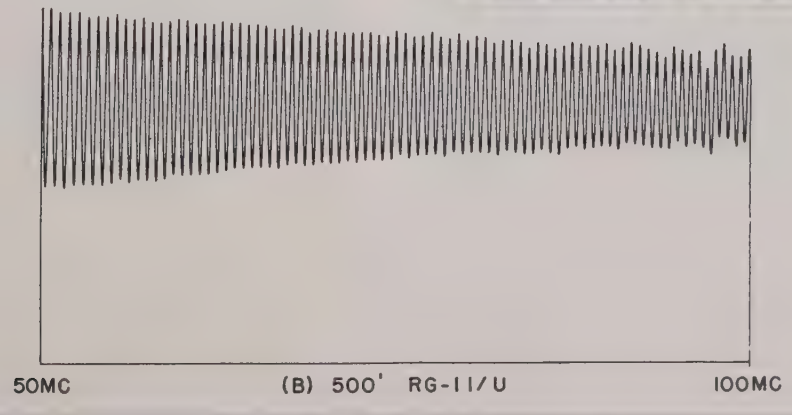


FIGURE 7

Terminations

Various Resistive Terminations

The wave reflected from a purely resistive termination is in phase with the main wave at the load for resistance values higher than Zo, or 180° out of phase for values lower than Zo, and the amplitude of the reflection increases as the resistance differs from Zo.

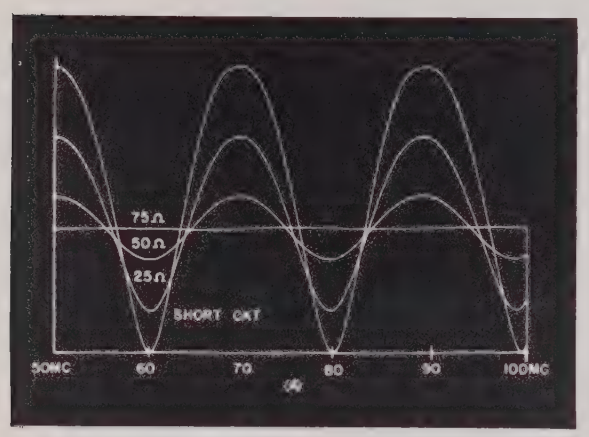


FIGURE 8a

Fig. 8 (a) shows superimposed the ripple patterns obtained with a low-loss 20 mc delay line, a sweep from 50 to 100 mc, and the indicated terminations. Note that all the patterns have minima at the same frequency as that with a shorted end.

> The impedance of a load giving a minimum at the same frequency as a short circuit is resistive and lower in resistance than Zo.

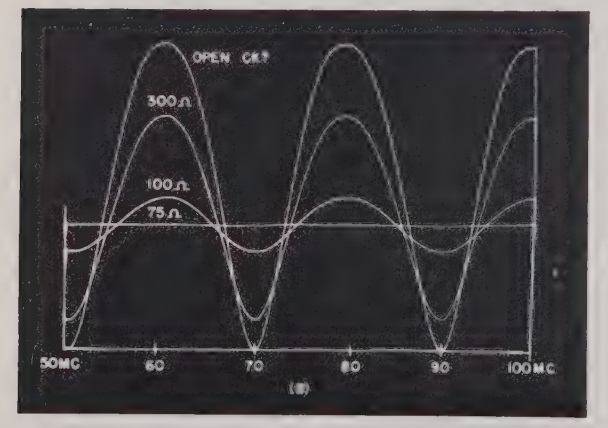


FIGURE 8b

Figure 8 (b) shows the patterns from several resistive terminations higher than Zo. They all have maxima at the "short-minimum" frequencies.

> The impedance of a load giving a maximum at the same frequency as a short-circuit minimum is resistive and higher in resistance than Zo.

Purely Reactive Terminations

A pure reactance does not dissipate energy, so the wave is reflected from a purely reactive termination at full amplitude. Its phase is shifted depending on the magnitude of the reactance relative to Zo. Thus the ripple pattern obtained with purely reactive terminations has the same amplitude as with an open or short circuit, but the minima and maxima are shifted in frequency. Fig. 9 (a) shows the patterns obtained with two sizes of capacitors compared with open and short circuit patterns.

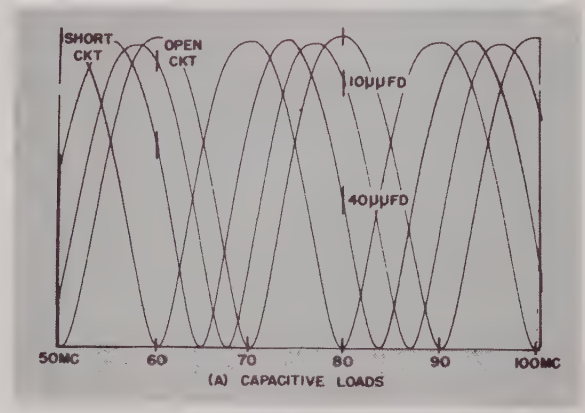


FIGURE 9a

The ripple pattern resulting from a capacitive termination has minima falling higher in frequency than those due to a short circuit, and lower than those due to an open circuit. This may be stated another way:

> If we mark the frequencies of short circuit minima (see marks on Fig. 9a) the marks will fall on down slopes with a capacitive load.

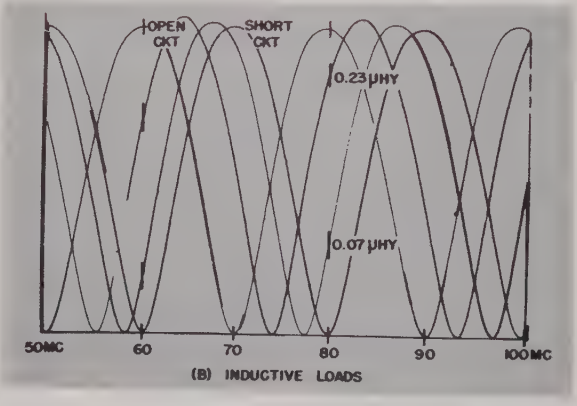


FIGURE 9b

Fig. 9 (b) compares the ripple patterns for two sizes of purely inductive terminations with those from an open and a short circuit.

> The ripple pattern from an inductive termination has minima falling above those from an open circuit, and below those from a short circuit. If we mark the frequencies of short circuit minima, the marks fall on upslopes for inductive terminations.

Complex Terminations; Impedance Changes with Frequency

When a termination has both dissipation (resistance) and reactance the reflected wave is reduced in amplitude, and shifted in phase relative to the main wave. Correspondingly, the ripple pattern has a lower amplitude, and minima shifted in position compared with a short or open circuit. Fig. 10 (A) illustrates the ripple pattern obtained with a low-loss 5 mc delay line and a series RC termination. At higher frequencies, where the reactance is lower, the termination approaches a matched condition, and the ripple has lower amplitude. The nature of the termination can be seen more clearly if the ripple pattern is compared with that obtained with a short circuit, and that obtained with a matching resistance. Fig. 10 (B) shows the three patterns superimposed. By observing the fact that down-slopes occur at shorted-minimum frequencies, and the ripple gets smaller at high frequencies, we could conclude that the load had the characteristics of a series RC circuit.

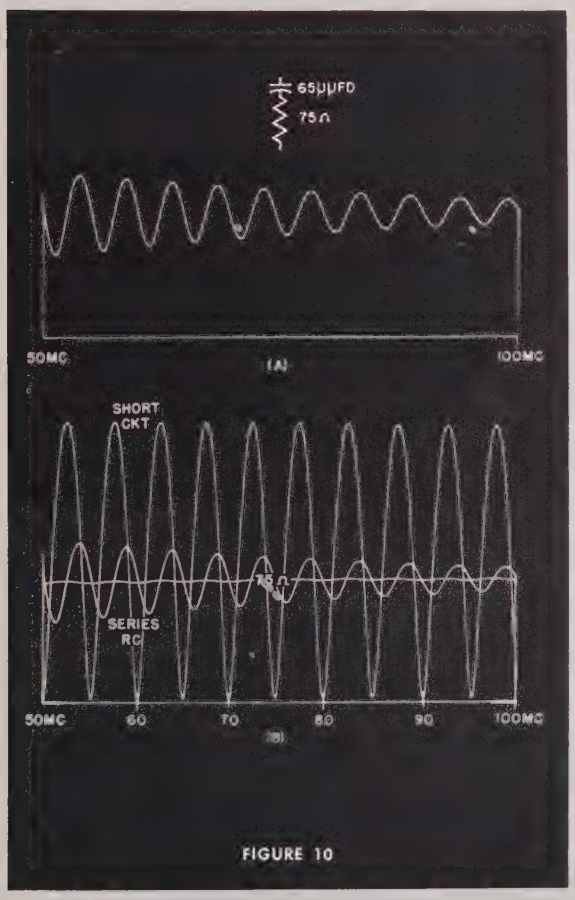


FIGURE 10

Fig. 11 (A) shows the ripple pattern with a 20 mc line and a series resonant circuit. Fig. 11 (B) shows the same pattern superimposed on a short circuit pattern and one from a matched resistor. By observing that its impedance is capacitive below the resonant frequency (down-slope at short-circuit minimum frequencies), matched at resonance (low ripple amplitude near 75 mc), and inductive above resonance (up-slope of shorted-minimum) we could deduce that it was a series RLC circuit with R equal to Zo.

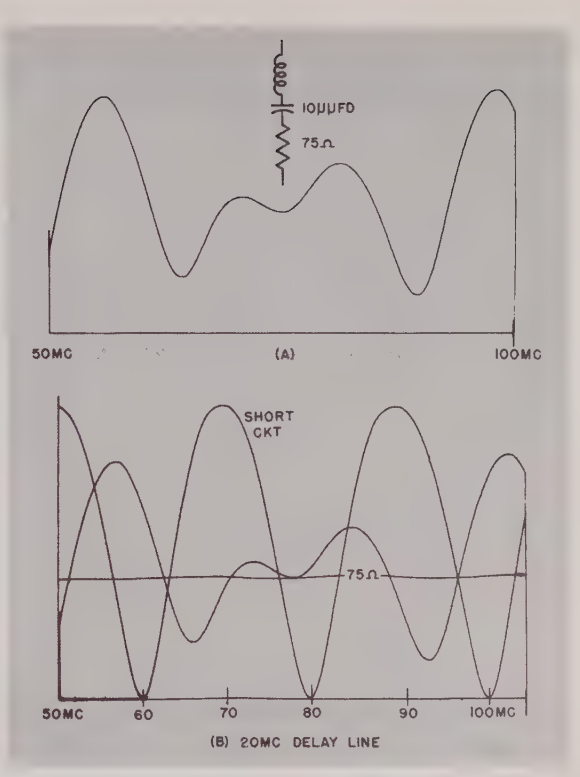


FIGURE 11

Fig. 12 (A) and (B) show the ripple patterns obtained under the same conditions as Fig. 11 (b) but with 5 mc and 2.5 mc delay lines respectively. They show how increased length in the delay line depicts the impedance characteristic in more detail.

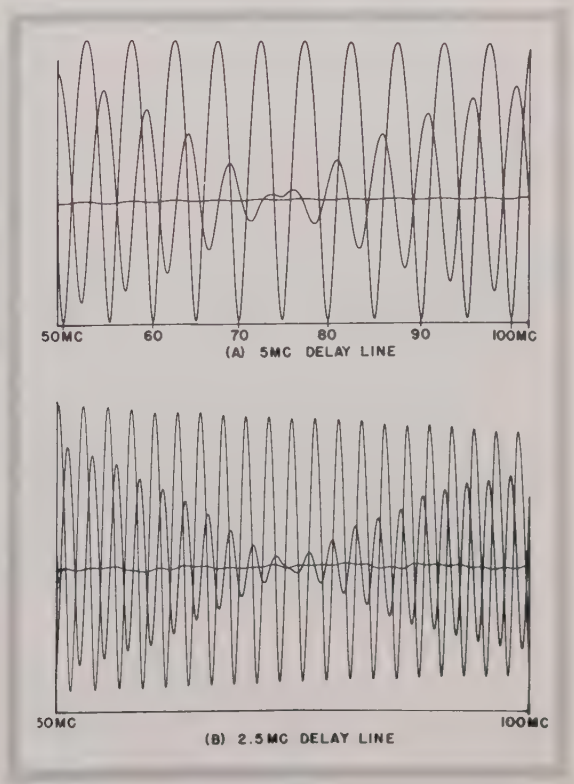


FIGURE 12

Determining Reflection Coefficient and VSWR

There is a temptation, in using the delay-line technique, to assume that the VSWR of the load is found by taking the ratio of minimum to maximum of the ripple pattern displayed on the scope. Two factors, the loss of the delay line and the non-linearity of the detector, make this procedure quite inaccurate. A more accurate procedure that eliminates the effect of the line loss and minimizes detector non-linearity is to compare the amplitude of the ripple pattern from the unknown with that from a short- or open-circuit.

Let A be the peak-to-peak amplitude of the ripple pattern obtained at the frequency of interest with the line shorted. Let B be the peak-to-peak amplitude of the ripple pattern from the unknown at this frequency.

Then the reflection coefficient of the unknown (ratio of main wave to reflected wave): K equals $\frac{B}{A}$

Percent reflection equals 100K

The *Return Loss* (reflection coefficient as a DB ratio) is $20 \log_{10} \frac{1}{K}$

The VSWR (ratio of max voltage at load to min. voltage is $\frac{1 + K}{1 - K}$ These various ways of expressing the degree of mismatch are conveniently related by the nomogram shown in Fig. 13.

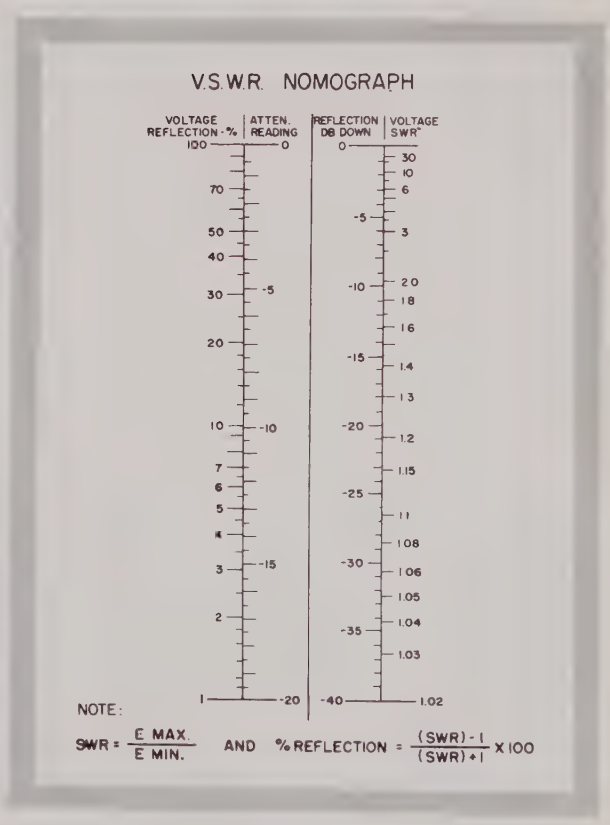

V.S.W.R. NOMOGRAPH
VOLTAGE REFLECTION - %
ATTEN. READING
REFLECTION DB DOWN
VOLTAGE SWR
NOTE:
SWR = E MAX. / E MIN. AND % REFLECTION = (SWR) - 1 / (SWR) + 1 X 100

FIGURE 13

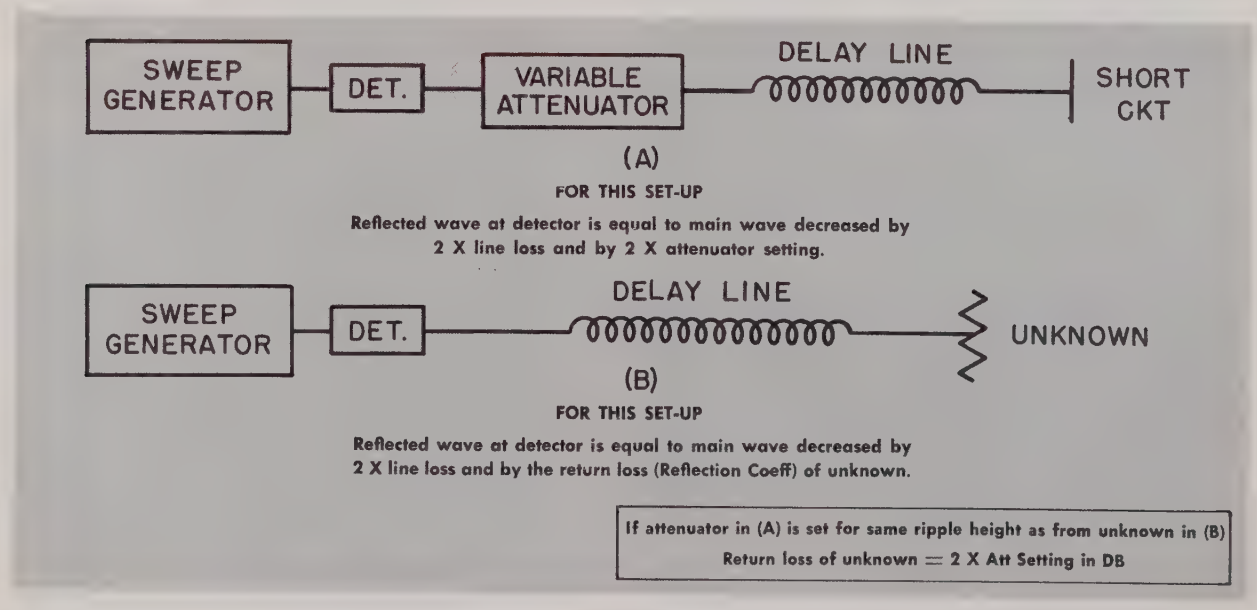

SWEEP GENERATOR
DET.
VARIABLE ATTENUATOR
DELAY LINE
SHORT CKT
(A)
FOR THIS SET-UP
Reflected wave at detector is equal to main wave decreased by 2 X line loss and by 2 X attenuator setting.
SWEEP GENERATOR
DET.
DELAY LINE
UNKNOWN
(B)
FOR THIS SET-UP
Reflected wave at detector is equal to main wave decreased by 2 X line loss and by the return loss (Reflection Coeff) of unknown.
If attenuator in (A) is set for same ripple height as from unknown in (B)
Return loss of unknown = 2 X Att Setting in DB

FIGURE 14

A more accurate, and generally more convenient way of determining the reflection coefficient of a load is to compare the height of its ripple pattern with the height of the pattern obtained with a short at the far end of the line, and a variable attenuator inserted between the line and the detector. Fig 14 illustrates the calibration set-up. By putting the attenuator at the detector end of the line the ripple height is determined primarily by its attenuation, and its impedance match is less important than if it were connected at the load end of the delay line.

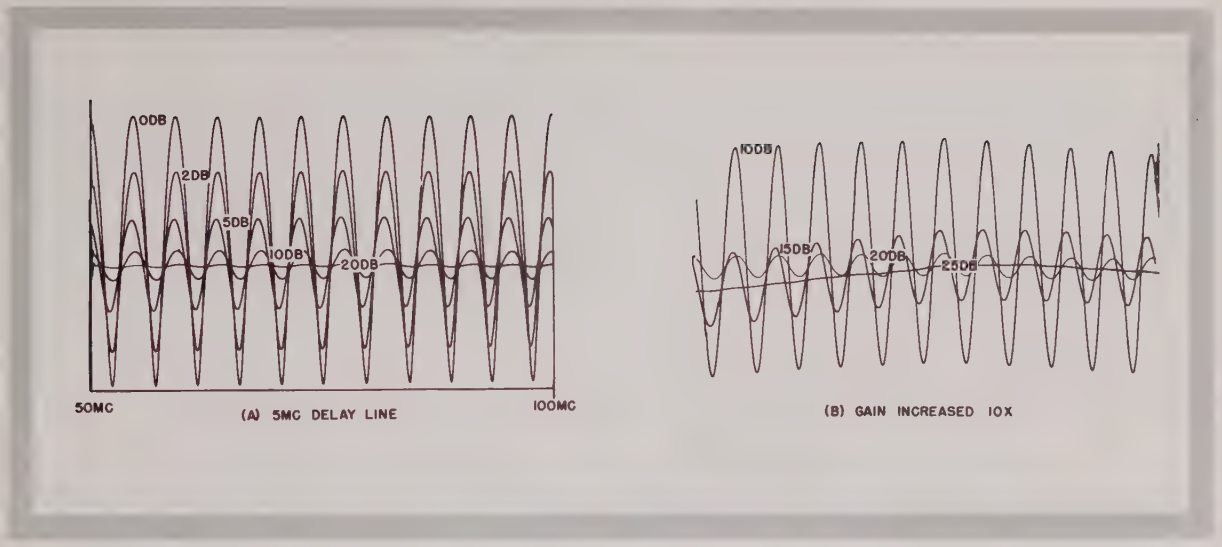


FIGURE 15

Fig. 15 illustrates the ripple patterns obtained at various settings of an attenuator connected in this way. Note that the ripple height with, for example, 5 db set on the attenuator, is what would be seen from a load having a return loss of twice this many db (e. g. 10 db). Fig. 15 (B) was made by increasing the vertical gain 10X compared with (A). Note that the ripple corresponding to a return loss of 40 db (VSWR 1.02) is readily seen. Also, note the irregularities in the way these traces match up. These are due to very small errors in the impedance match and delay of the attenuator used.

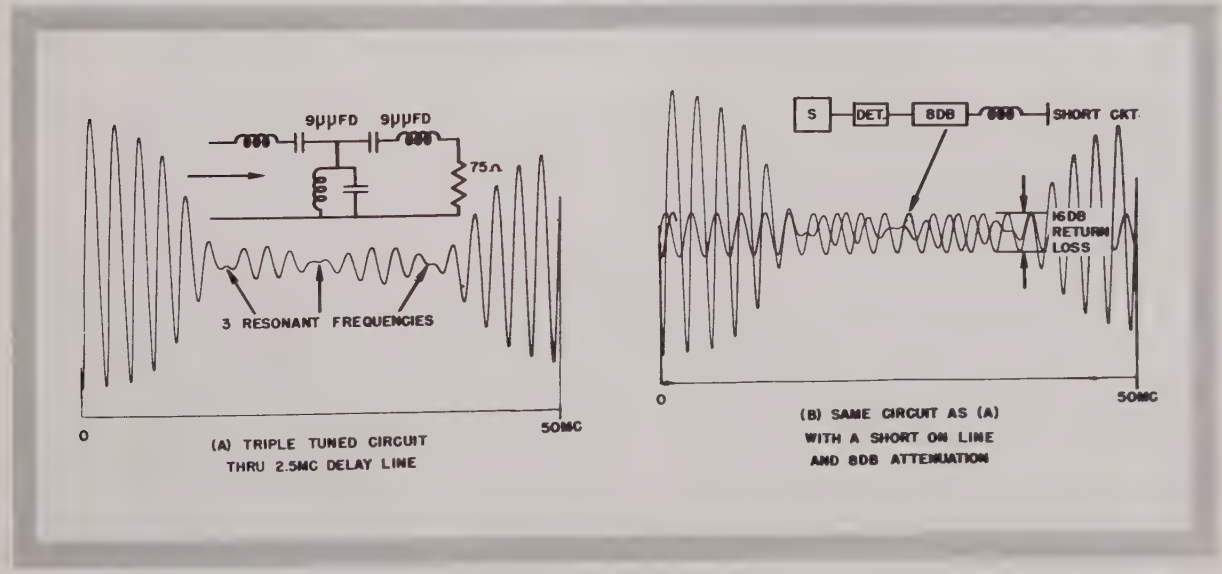


FIGURE 16

Fig. 16 illustrates how the technique is used in determining return loss. A triple-tuned band-pass filter was connected to the end of a 2.5 mc delay line, giving the ripple pattern of Fig. 16 (A). Note the three frequencies of best match. To determine the maximum reflection within the pass band, the ripple from a shorted line through a variable attenuator was set to the same height as the maximum ripple in the pass band. The attenuator read 8 db. indicating a maximum return loss for this filter of 16 db. The two patterns are shown superimposed in Fig. 16 (B).

Irregularities in the Delay Line

With reasonable sweep output level and scope gain, it is not difficult to see ripples corresponding to 40 or even 50 db return loss. It is not generally possible to make measurements with this much accuracy. The limiting factor is not gain, but the uniformity of the delay time. Many commercial coaxial cables have a degree of nonuniformity that results in appreciable reflections from within the cable, even when terminated with the best possible load. A poor cable may have reflections with a return loss of as little as 20 db. Most cables run at or above 30 db., and only an exceptionally uniform cable has internal reflections more than 40 db down. To illustrate the extremes that may be encountered several traces were made. Fig. 17 (A) shows the ripple pattern of a 5 mc delay line of exceptional uniformity terminated in a load that closely matches its impedance. It was possible to see variations in the pattern only by increasing the vertical gain 10X (Fig. 17B). Comparing the ripple pattern with the calibration pattern corresponding to 40 db return loss, it can be seen that the combined return loss of cable and termination is decidedly better than 40 db over the whole frequency range of the sweep. For contrast, Fig. 17 (C) shows the trace with a very poor piece of cable. Even with the best termination possible, its variation is more than 10X greater than the other cable. (compare with 17A).

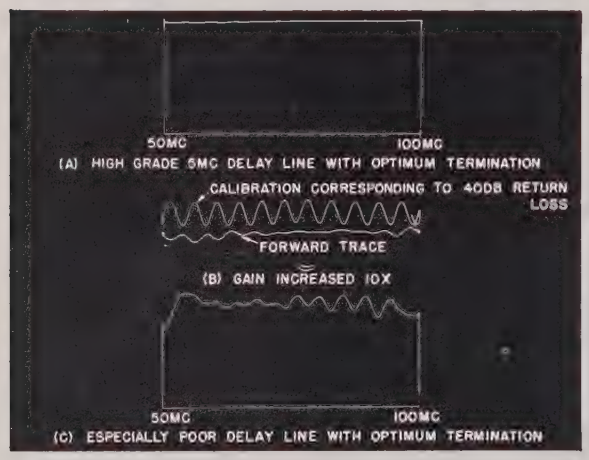
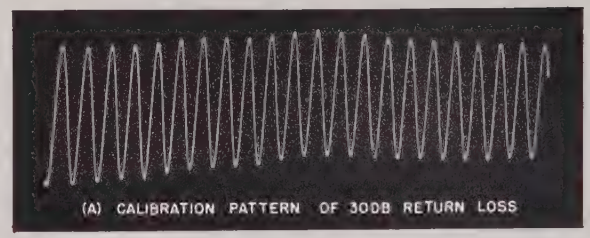


FIGURE 17



(B) OPTIMUM TERMINATION

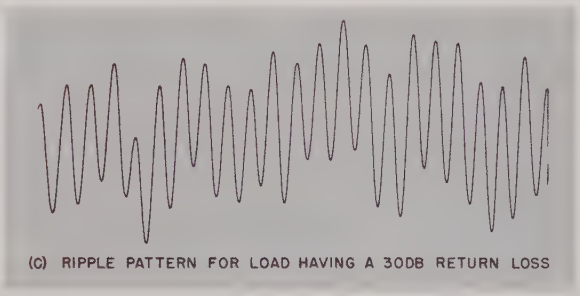


FIGURE 18

Fig. 18 is an illustration of the way in which irregularities in the delay line can make accurate measurement difficult. (A) shows the ripple pattern amplitude under a calibration condition corresponding to 30 db return loss (high vertical gain was used). (B) shows the pattern with termination adjusted for minimum ripple. This line, a 150' length of RG11/U, has internal reflections a little more than 30 db down. (C) shows the pattern resulting when a load having a return loss of 30 db was connected to the end of this line. It can be seen that the line irregularities prevent accurate display of the load characteristics. A further hazard is that the internal reflections probably indicate a considerable variation in characteristic impedance of one section of the cable as compared with another. Thus, the impedance which will match the far end depends on just where the cable is cut.

A Note on Comparison Technique

The patterns illustrating this article were recorded with a Moseley X-Y recorder and sweep sampling attachment. A Jerrold Model 900B Precision Sweep Generator was used in these tests. Where patterns are shown superimposed, they were made by simply changing the load connections without touching the sweep settings and recording the second trace on top of the first. This is, of course, not possible with a scope, but the same convenience of simultaneous presentation for calibration can be obtained by using a Jerrold Model FD-30 or TC-3 Coaxial Switch. For Delay Line work, it is connected as shown in Fig. 19, superimposing a calibration pattern from a second delay line on the pattern of the load being adjusted or measured.

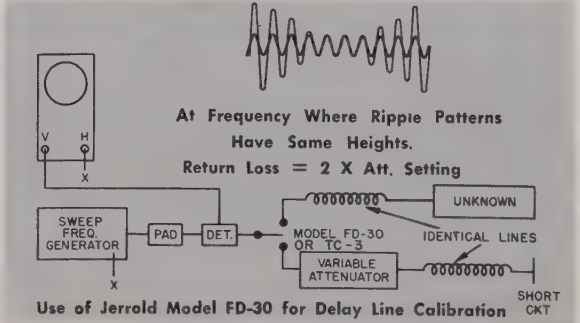


FIGURE 19

NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila., Pa. 19132. Phone: BAldwin 6-3456 — Ext. 215.

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COMPARISON TECHNIQUE
IMPROVES SWEEP FREQUENCY MEASUREMENTS*

By K. A. SIMONS, Chief Engineer, Jerrold Electronics Corporation

The Jerrold Electronics Corporation maintains a Design Engineering Laboratory at Hatboro, Pennsylvania. The primary function of this organization is the design of electronic equipment, including wide-band amplifiers, filters and associated equipment in the "awkward" frequency range between 5 and 220 mc. This range is called awkward because the upper end is too high for measuring methods suitable for the low communications frequencies, and the lower end is too high for standing-wave line measurements. A number of measuring techniques, developed to meet this situation, are described in this article.

*A major part of this article appeared first under the title "SWEEP MEASUREMENTS MAKING THE TOUGH ONES EASY" in "electronics" Vol. 36, No. 16, April 19, 1963.

BASIC COMPARISON CIRCUIT

The basic technique is to compare transmission through two paths. Figure 1 shows the method used to determine accurately the transmission loss through a network. A test signal source and an output indicator are connected to a DPDT switch so that in one switch position (TEST) the signal is passed through the network the loss of which is to be measured, and in the other switch position (REFERENCE) it is connected through a variable calibrated attenuator.

To make a measurement, the operator throws the switch repeatedly back and forth, while adjusting the attenuator until the output indicator shows identical responses for both switch positions. The loss in the tested network is then equal to that of the attenuator and can be read off from its calibration.

This technique has been used for many years for precise transmission measurements in the telephone art. (See, for example, the article "Transmission Measuring for Manufacture," by George W. Pentico in THE WESTERN ELECTRIC ENGINEER, October, 1961.)

The technique has some of the basic advantages of the more familiar null-bridge measuring method, namely:

1. High source output and a sensitive indicator make very precise settings possible.
2. Other characteristics of the source and indicator have little effect on the accuracy of the result. Neither needs to be calibrated, and the amplitude stability need only be constant long enough for a comparison to be made.
3. Calibration accuracy is entirely determined by a convenient passive attenuator.

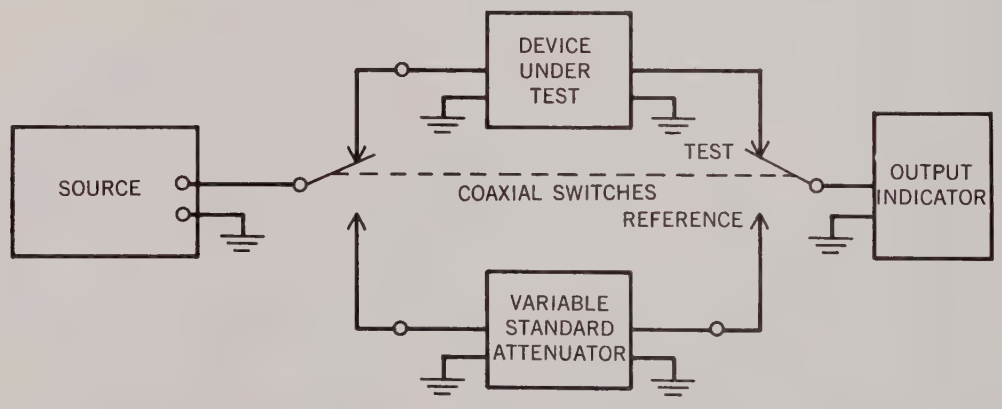


Fig. 1

COMPARISON WITH SWEEP DISPLAY

This approach is a "natural" for sweep frequency testing over wide frequency bands. Sweep frequency generators and oscilloscopes are notoriously as poor on calibration as they are good on sensitivity, while stable and accurate attenuators are common. To adapt to the sweep situation, the basic technique is modified in two respects:

1. High frequency measurements necessarily involve coaxial transmission lines; thus the comparison switch, as part of the transmission system, must be coaxial with an impedance matching that of the associated lines.
2. Oscilloscopic display of a sweep frequency response curve requires periodic repetition of the measurement (usually 60 times per second) so the eye can see a steady pattern. The "switch" must also operate period-

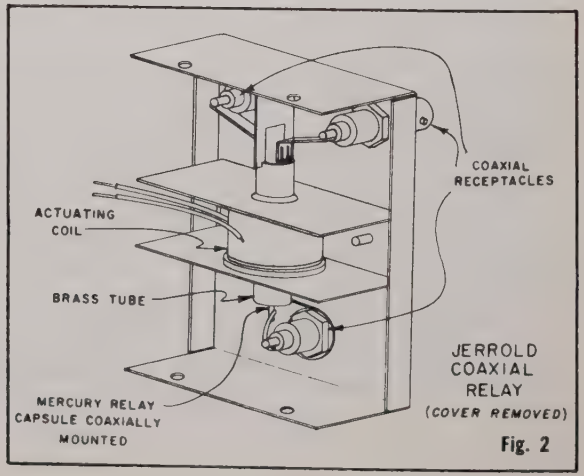


Fig. 2

ically so that the calibration may be seen as part of this stationary pattern. To achieve periodic switch operation, a high speed coaxial relay is used. The construction is illustrated in Figure 2. A Clare mercury-wetted relay capsule is mounted so that it forms a well-matched coaxial structure. This unit operates at high speeds (up to 60 cps) with no bounce and with very long life.

Switches have been in full-time operation in our laboratory for five years with negligible failures.

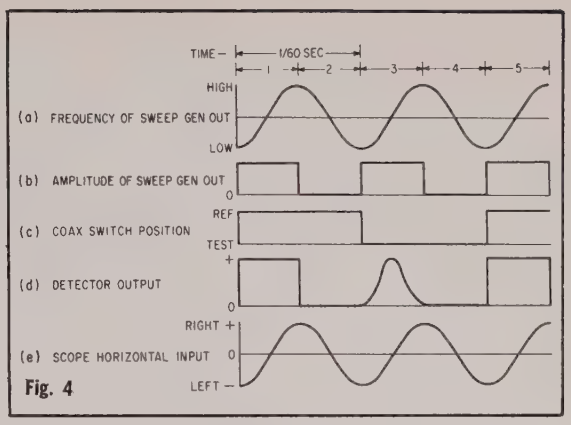


Fig. 4

MEASURING LOSS BY COMPARISON

The test set-up for measuring and displaying the high frequency loss of a network by the sweep-frequency comparison method is diagrammed in Figure 3. The output of a Sweep Frequency Generator, sweeping across the desired band at a 60-cycle rate, is connected to one of two SPDT coaxial switches. The other switch is connected to the input of a wide-band detector. The switches transmit alternately through the network under test and through the variable attenuator.

When this detector output is applied to an oscilloscope's vertical input terminals, and a voltage, with its instantaneous magnitude phased to match the instantaneous frequency of the generator, is applied to the horizontal output terminals, the scope traces out the display illustrated in Figure 4f. (In the example shown, the horizontal deflecting voltage must be a 60-cycle sine-wave phased as shown, to agree with the waveform and phase of the frequency modulation.)

The curve presents the loss as it varies with frequency in the unit under test, together with a base line showing zero response (infinite loss) and the REFERENCE trace showing the loss corresponding to the attenuator setting.

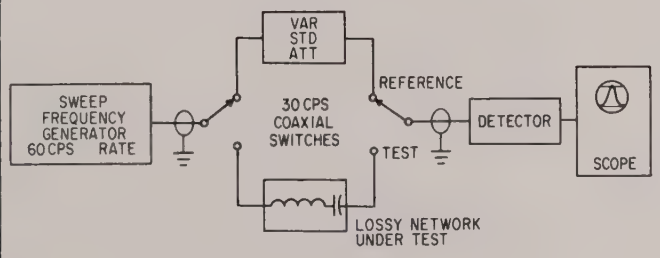


Fig. 3

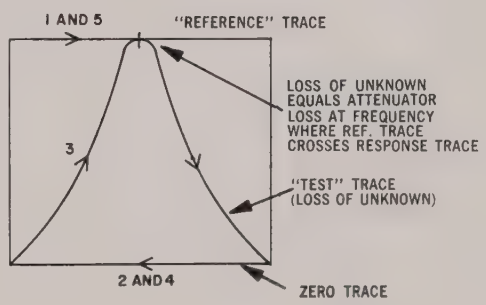

1 AND 5
"REFERENCE" TRACE
LOSS OF UNKNOWN EQUALS ATTENUATOR LOSS AT FREQUENCY WHERE REF. TRACE CROSSES RESPONSE TRACE
3
"TEST" TRACE (LOSS OF UNKNOWN)
2 AND 4
ZERO TRACE
Fig. 4f

Figure 4 illustrates the manner in which the display is generated. The frequency of the sweep output varies sinusoidally between an upper and a lower limit (Figure 4a) at a 60-cycle rate. The output is cut off during the return trace (Figure 4b) to form a zero reference trace. The coaxial switches are transferred at a 30-cycle rate (Figure 4c) allowing the development of a full trace in the REFERENCE position followed by a full trace in the TEST position. As a result, the detector output (Figure 4d) goes through four conditions in each thirtieth of a second; first indicating reference output, second delivering zero output, third following the loss curve of the unit under test and fourth repeating the zero output trace.

The REFERENCE trace crosses the TEST trace at those frequencies where the loss of the unit under test precisely equals the loss of the standard attenuator.

The scope trace photograph in Figure 5 shows the insertion loss measurement of a triple-tuned band-pass filter. The attenuator was set at 1.0 db and the point at which the REFERENCE trace crosses the response shows a loss of 1.0 db at a 7-mc band-width centered on 125 mc. It is a major advantage of this presentation that the accuracy of the measurement is not affected by the constancy of the sweep output across the band, the flatness of the detector, or by drift or other changes in scope gain and centering.

In cases where very small attenuation differences are to be observed, it is convenient to increase the vertical gain and recenter the trace to show a highly expanded view of part of a response pattern. This is illustrated in the series beginning with Figure 6. A Jerrold 900-B sweep frequency generator was set to sweep between zero and 200 mc, and a short coaxial jumper was connected between its output and a detector. The scope vertical gain was increased to about four times that required to show the change from zero to full output, and the centering readjusted to bring the resulting trace onto the screen.

Under this condition the screen showed a change of only 0.2 db per division and the resulting trace (Figure 6) shows the sweep output to be constant within ± 0.2 db. With the same sweep and scope settings, the sweep output was then connected through two coaxial switches (Jerrold Model FD-30) transferring between a short jumper and a variable attenuator (Jerrold Model AV-50) connected through two 2-foot coaxial cables.

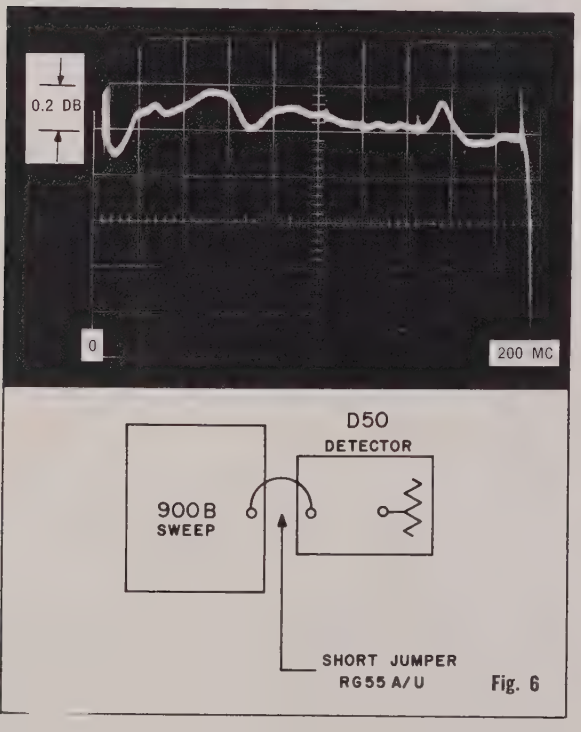
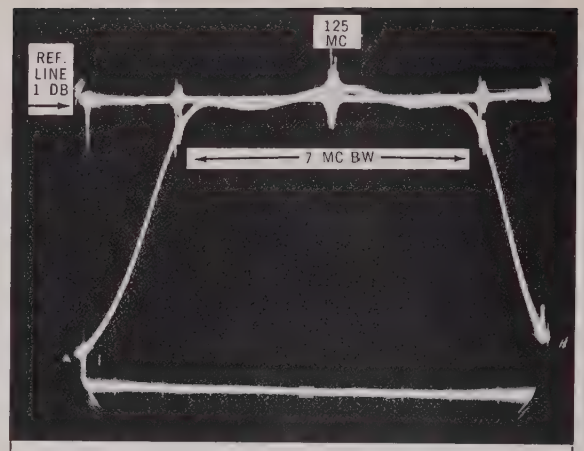


Fig. 6



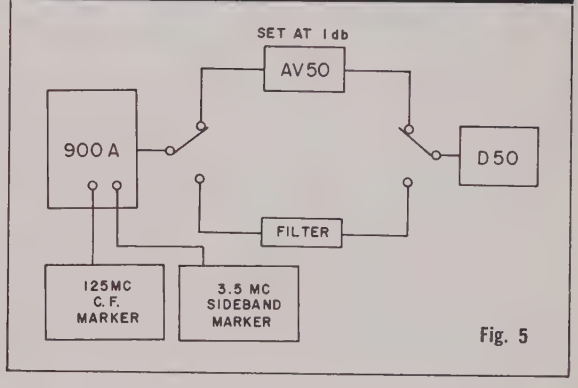
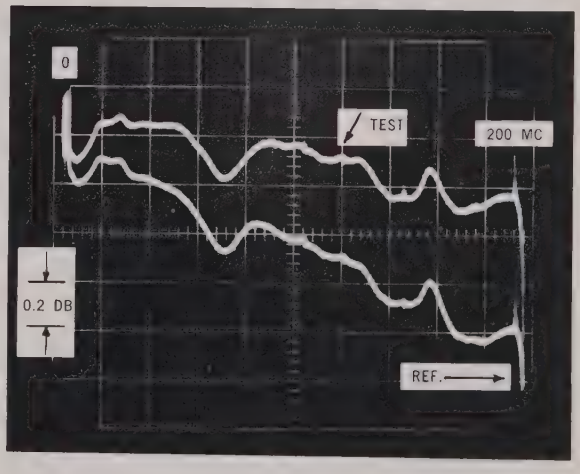


Fig. 5



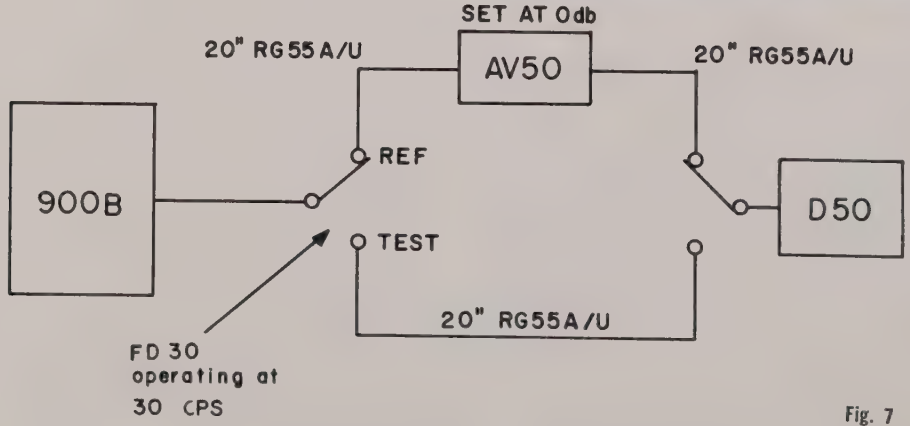


Fig. 7

The resulting photograph (Figure 7) compares the jumper loss with the attenuator insertion loss (at zero setting) continuing with the same expanded vertical scale. By careful inspection it can be seen that the TEST (upper) trace is slightly down at the high end (compared with Figure 6) indicating that the coaxial switches and short jumper have added about 0.2 db loss at 200 mc, while the REFERENCE (lower) trace shows that the attenuator and its longer cables have added about 0.8 db loss at the high end.

If precise measurements are to be made it is **not** essential that either curve be precisely flat, but it **is** essential that the loss on the TEST side, before inserting the unknown, be the same as that on the REFERENCE side. This condition was satisfied by inserting additional cable on the TEST side (totalling 137 inches) so that the insertion losses in both circuits were close to identical. The resulting trace was photographed (Figure 8a) and shows considerably less than 0.05 db difference between the two circuits over the whole frequency range from low frequencies up to 200 mc. Inserting an additional 0.5 db into the standard attenuator gave the result illustrated in Figure 8b, showing that this set-up could be used for loss measurements involving very small attenuation differences.

Figure 9 shows how this approach was applied to a practical measuring problem. We wanted to test a repeater amplifier designed for long cascaded runs in television distribution systems. The amplifier, together with its equalizer, is required to have gain equalizing as closely as possible the loss of 2000 feet of a particular coaxial cable between the frequencies of 5 and 95 mc. To test this, the cable, the equalizers and the amplifier were connected, each following the other on the TEST side of the comparison circuit, with a jumper providing a zero loss reference on the REFERENCE side. The resulting response (illustrated in Figure 9b) shows that the combination of cable plus equalizer plus amplifier had zero loss within about $\pm$ 0.25 db between 5 and 95 mc. The fine grained ripple is primarily due to variations in the cable's transmission characteristic.

The usefulness of this technique is obvious, first in the fact that it would take many hours to make this measurement using point-by-point techniques and even then the accuracy would be questionable, and second in the fact that the necessary refined adjustments of the equalizer and amplifier can be carried out quickly and accurately while observing the display, which could not be done with a point-by-point approach.

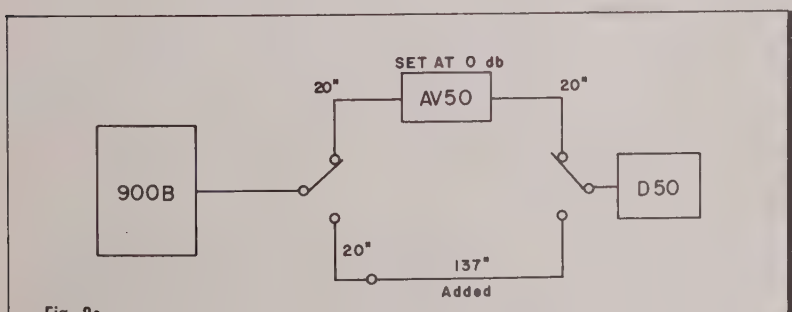


Fig. 8a

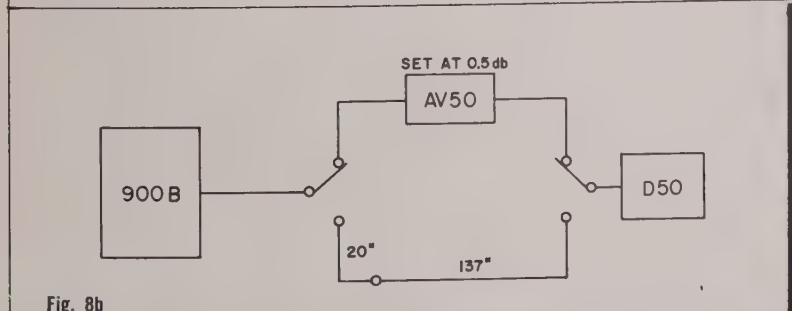


Fig. 8b

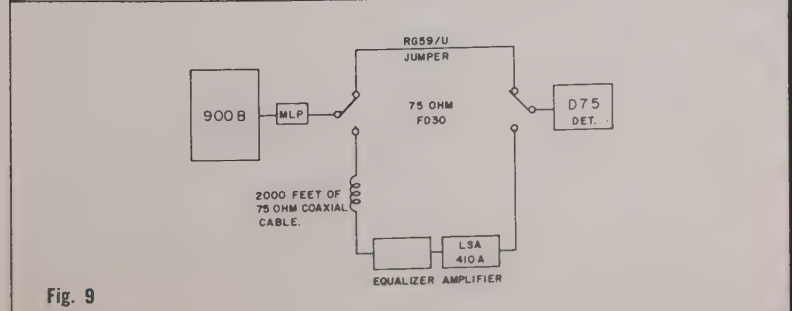
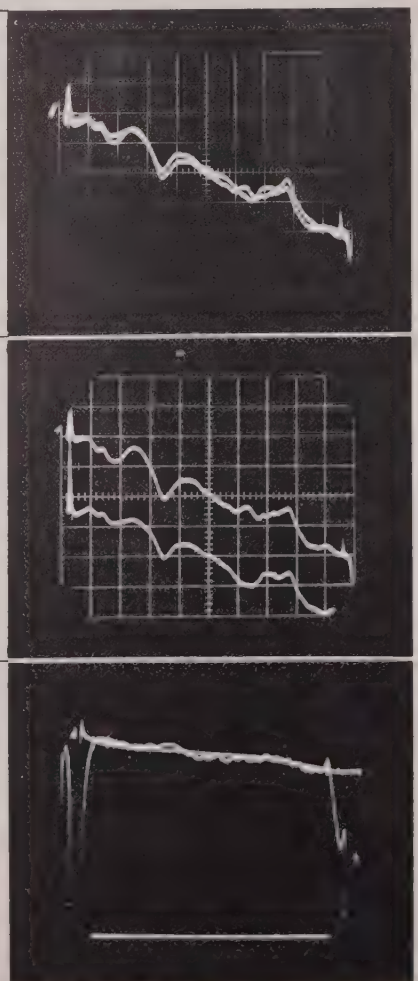


Fig. 9



The comparison technique can be used effectively for the measurement of amplifier gain. The basic circuit is diagrammed in Figure 10. By using a jumper as a zero loss reference and inserting the variable attenuator ahead of the amplifier on the TEST side, the REFERENCE line is made to cross the amplifier response curve at the point where the attenuator loss equals the amplifier gain. The resulting display is illustrated in Figure 11.

The photographs in Figure 12 illustrate gain measurement on an amplifier designed to pass television channel 13 for use in rf distribution systems. Figure 12a illustrates the response curve in the TEST position with a frequency marker at band center (213 mc). Figure 12b illustrates a useful frequency marking technique where a low frequency generator (3 mc) is connected simultaneously with a high frequency one and the resulting side-bands mark the channel edges (213 ± 3 mc). Figure 12c illustrates the trace with the coaxial switch running and the attenuator set at 57 db (0.5 db below the maximum amplifier gain).

To obtain a "blown-up" view of the pass band response, the scope gain was turned up and the trace recentered to give the result shown in Figure 12d. This shows that the amplifier response is within 0.1 db across the band. Neither the measurement, nor the alignment which preceded it, could be readily carried out by any single frequency measurement technique.

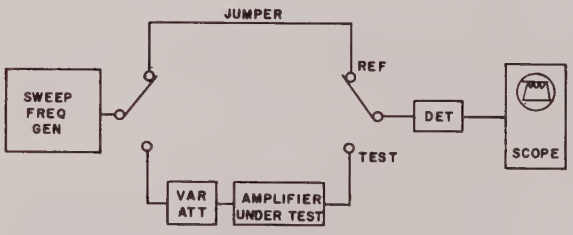


Fig. 10

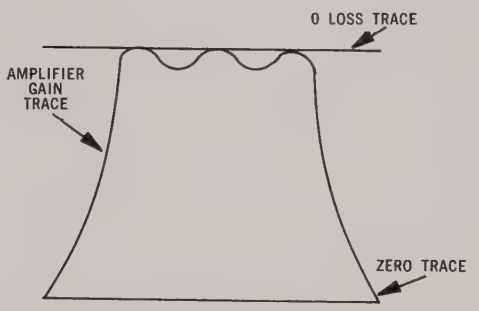


Fig. 11

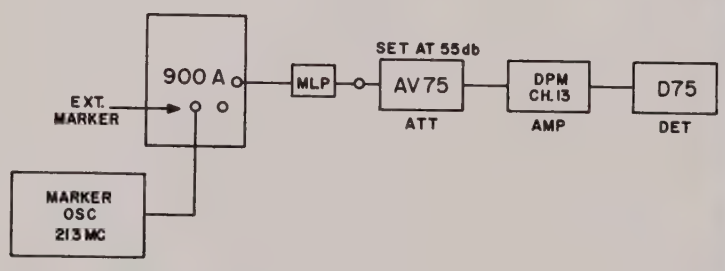
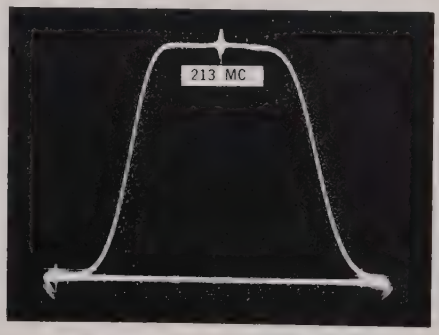


Fig. 12a



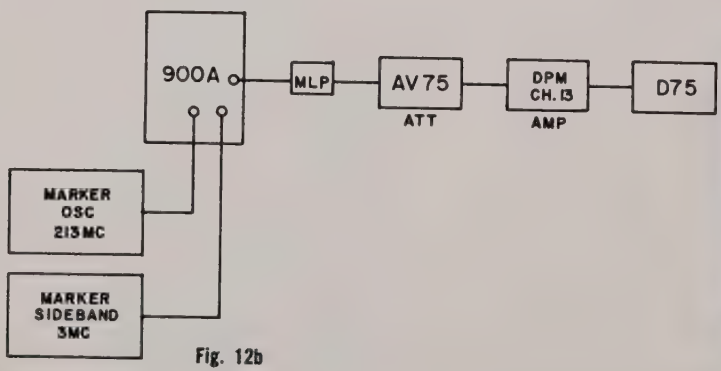
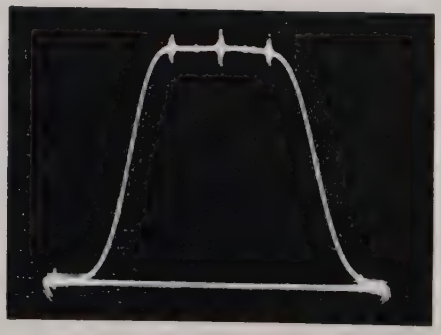


Fig. 12b



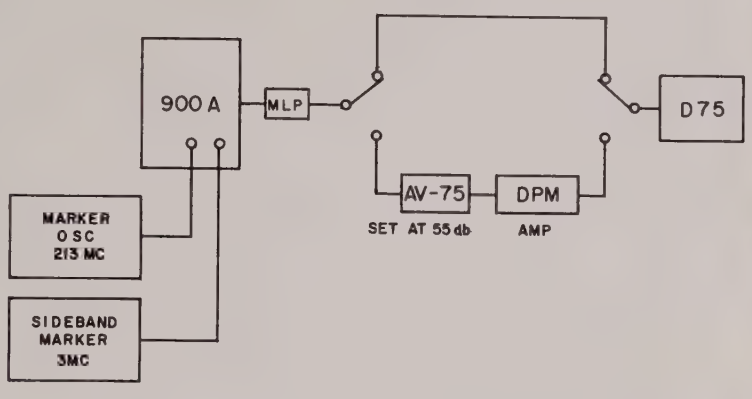
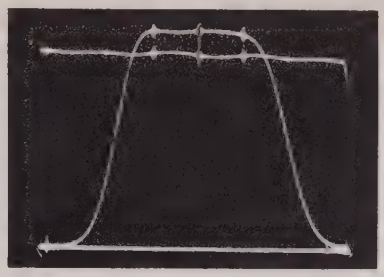


Fig. 12c



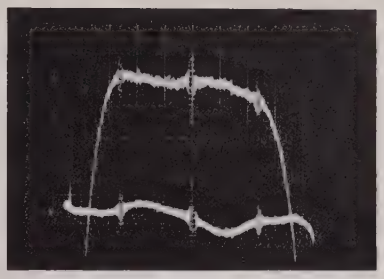


Fig. 12d

MEASURING IMPEDANCE MATCH BY COMPARISON

In addition to gain and loss measurements, a third application of the comparison method was found to be very useful in our laboratory. This involves employing a bridge circuit to provide wide-band display of impedance match characteristics. As described more fully in TNL, Vol. 1, No. 4, the technique depends on the characteristics of a high frequency bridge having six equal arms (including the impedance of the source and detector). The bridge circuit is diagrammed in Figure 13 and its effect is summarized by the fact that the insertion loss between the input and output terminals is equal to 12 db plus the return loss of that which is connected to the unknown terminal (return loss is the ratio of the reflected wave in a coaxial system to the incident wave, expressed in db). Stating this another way: when a signal is introduced into the input terminal of the bridge, the output signal **is** the reflected wave from the unknown diminished by 12 db.

Thus the bridge makes possible impedance measurements over a wide frequency band, converting impedance variations into loss variations which are then accurately measured by the comparison technique. It is, in effect, a directional coupler giving flat response over a very wide band of frequencies in the "awkward" 5 to 200 mc range. The equipment set-up for impedance sweeping is diagrammed in Figure 14. The bridge is connected to the TEST terminals of the comparison switches and a fixed 12 db attenuator is connected with the standard attenuator in the REFERENCE leg. To obtain adequate gain for the measurement of small reflections, a wide-band distributed amplifier is connected ahead of the detector. The display obtained is shown in Figure 14b, the REFERENCE trace crossing the unknown at a return loss level corresponding to the attenuator setting. The bridge we are now using allows measurement of return loss in the range 0 to 40 db, with an accuracy within $\pm$ 1 db.

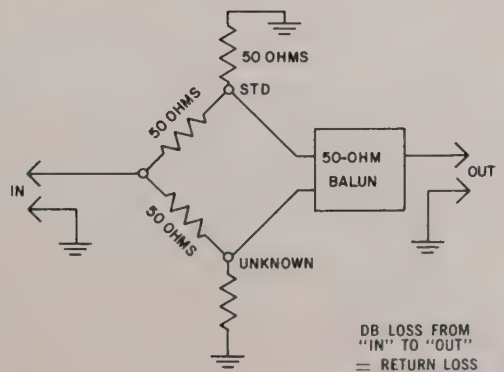


Fig. 13

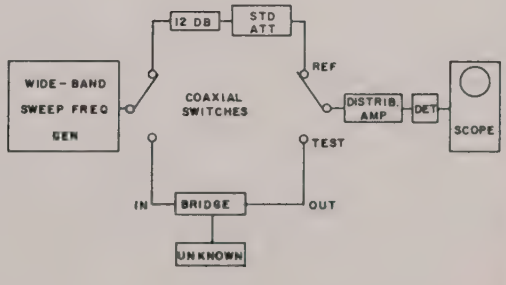


Fig. 14a

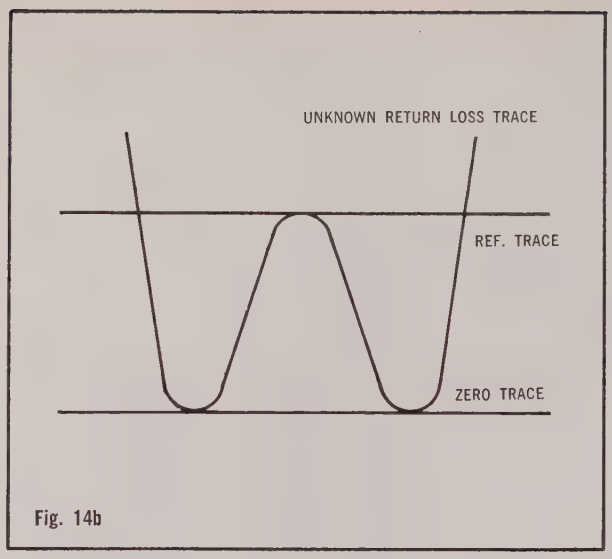


Fig. 14b

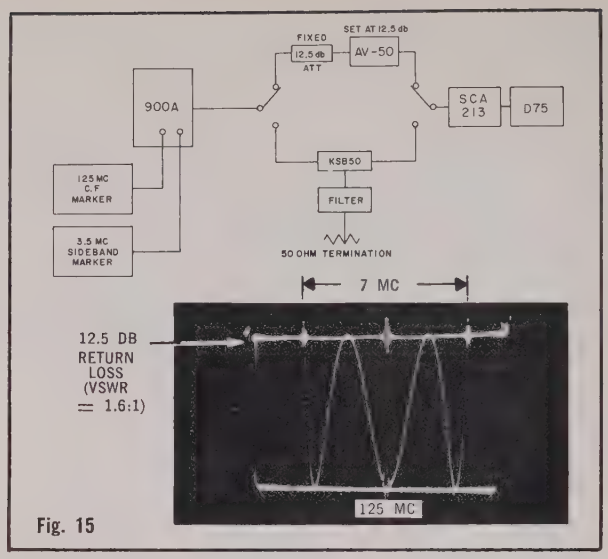


Fig. 15

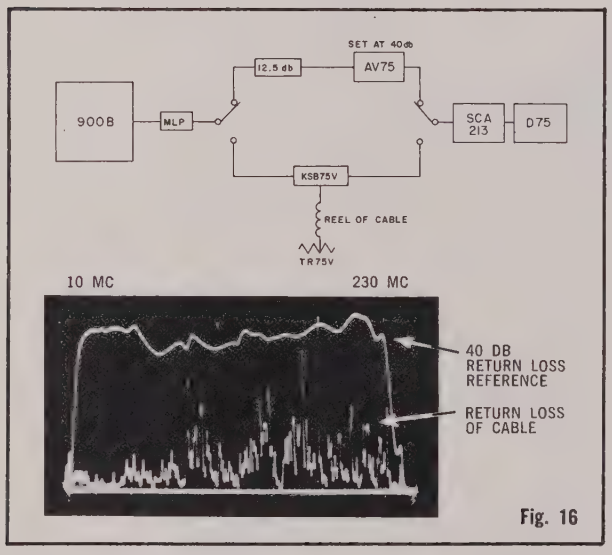


Fig. 16

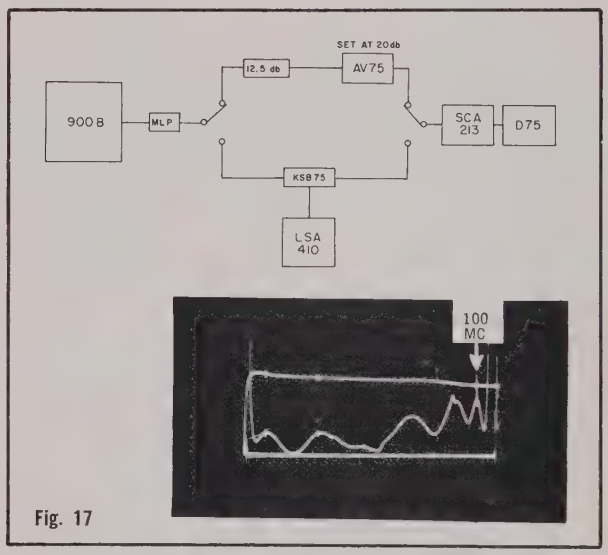


Fig. 17

Figure 15 illustrates measurement of the input impedance of the band-pass filter the insertion loss of which is shown in Figure 5. The filter was aligned for maximum return loss of 12.5 db (max. VSWR of 1.6:1) with Chebishev (equal ripple) match characteristics. Adjustment and measurement were carried out rapidly with the aid of the comparison presentation.

Figure 16 illustrates measurement of the "structural return loss" (the impedance variations due to minor mechanical irregularities) in a 1000-ft. piece of unusually good coaxial cable. The maximum reflection spike at about 150 mc had a return loss level just over 40 db.

Measurement of such a complex characteristic over this frequency range would be a maddening (if not impossible) process by any point-by-point technique.

Figure 17 illustrates the impedance match at the output terminal of the amplifier the response of which was shown in Figure 9. Without the rapid and accurate measurement of match made possible by the return loss bridge, production alignment of this amplifier to meet the required overall performance would be impossible.

CONCLUSION

The comparison technique has been described as it is used at the Jerrold Laboratory for measurement of loss, gain, and impedance match, simultaneously with sweep frequency display of these characteristics. This technique has been found absolutely indispensable in the speed and convenience it contributes to the design and adjustment of a variety of devices in the "awkward" frequency range between 5 and 220 mc.

NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila., Pa. 19132. Phone: BAldwin 6-3456 — Ext. 215.

JERROLD

Technical Newsletter

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NEW ATTENUATOR AND COMPARATOR

EXTEND FREQUENCY RANGE OF COMPARISON TECHNIQUE TO 1200 MC.

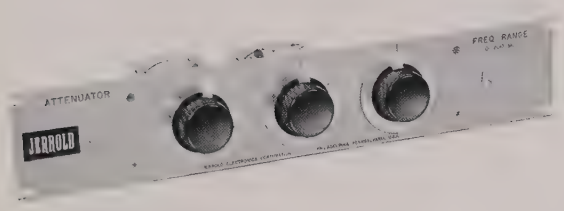


Fig. 2—Model ATV-109R Rack-Mounting Bank of Turret Attenuators

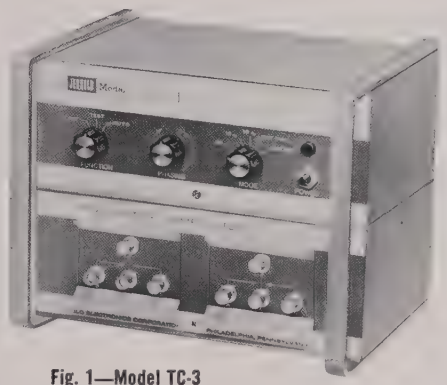


Fig. 1—Model TC-3 Coaxial Comparator

SUMMARY: A new drum attenuator and a transistorized three-way coaxial comparator are described, which allow accurate sweep-frequency measurements of transmission loss, gain, and reflection coefficient in 50-ohm coaxial systems from DC to 1200 MC.

INTRODUCTION

For the past six years, Jerrold has been offering as accessories to the 900-A and 900-B Sweep-Frequency Generators a comparator (high speed coaxial switch Model FD-30) and variable attenuators (Model AV-50). When used with a wide-band sweep-frequency generator these allow quite accurate measurements of transmission gain and loss[1], and when used with a suitable bridge, measurement of reflection coefficient (return loss)[2] to a limit of accuracy determined by the characteristics of these accessories. The FD-30 Comparator and the AV-50 Attenuator were originally designed for use in the frequency-range up to 250 MC, where they had excellent characteristics, with reasonably good performance up to 500 MC. Since the Model 900 Sweep Generators provide coverage up to 1200 MC there has been a need for improved accessories, making the comparison technique available up to the limit of the sweep capability.

The new Model TC-3 Transistorized Comparator and the Model ATV Turret Attenuators have been developed to meet this need. They provide measurement accuracy up to 1200 MC, somewhat better than that obtained with the older equipment up to 250 MC, and in addition feature several marked improvements in operating convenience over the older equipment.

1. Jerrold Technical Newsletter Volume 2, Number 1—March 1964. COMPARISON TECHNIQUE IMPROVES SWEEP FREQUENCY MEASUREMENTS. Ken Simons, Jerrold Electronics Corporation.

2. Jerrold Technical Newsletter Volume 1, Number 4—August 1960. A BRIDGE METHOD OF SWEEP FREQUENCY IMPEDANCE MEASUREMENT. (Same author.)

Figure 1 shows the Model TC-3 Coaxial Comparator. It consists of two basic housings arranged so that they can be mounted one on top of the other to form a convenient package for bench-mounting, or mounted side-by-side to form a 3½″ high panel in a standard 19″ rack. One of the units houses the transistor driver and the controls; the other houses the two coaxial switch assemblies. Each switch assembly consists of three reed relays mounted to form a single-pole triple-throw coaxial switch.

At frequencies in the gigacycle region, jumper cables present a major limitation on the accuracy of measurements. Therefore, accuracy is greatly improved if jumper lengths are kept to an absolute minimum. For this reason each assembly switch has been designed so that it can be removed from the housing and extended on a long cable so that it can be placed immediately adjacent to the units being tested. The reed relays operate equally well in any position; thus the switching heads can be positioned to suit the physical position of the units under test.

The operation of the new comparator can best be explained by comparing it with the FD-30. Figure 3a illustrates application of the FD-30 to the measurement and display of the frequency response of a resonant network. The sequence of operations when the unit is used with a 60-cycle supply is illustrated in Figure 3b and the resulting scope trace is shown in Figure 3c. Note that the complete cycle occurs in one-thirtieth of a second, and that the zero line is traced out twice in each cycle.

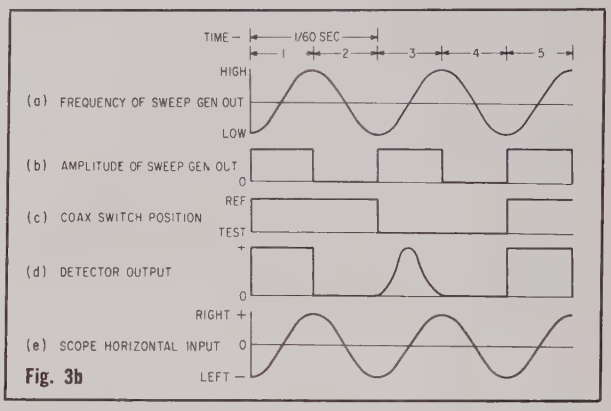


Fig. 3b

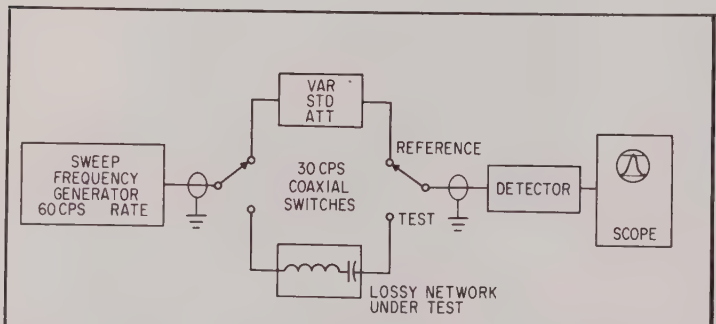


Fig. 3a—Operation of 2-Way Comparator Model FD-30

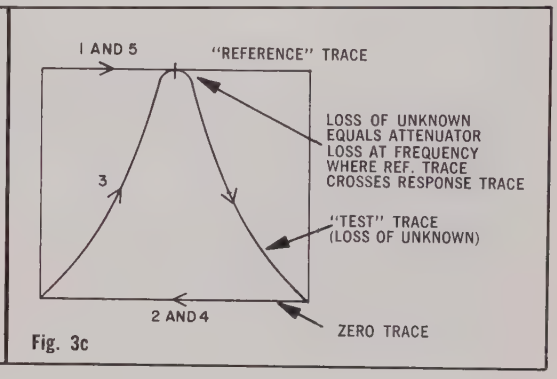

1 AND 5
"REFERENCE" TRACE
LOSS OF UNKNOWN EQUALS ATTENUATOR LOSS AT FREQUENCY WHERE REF. TRACE CROSSES RESPONSE TRACE
3
"TEST" TRACE (LOSS OF UNKNOWN)
2 AND 4
ZERO TRACE

Fig. 3c

Figure 4a illustrates the application of the three-way comparator using two attenuators to show simultaneously the maximum response and the 3 db down point of a resonant circuit on an oscilloscope. Figure 4b diagrams the sequence of operations. Note that the switches "A" close during the first quarter cycle, connecting attenuator "A" into the circuit; switches "B" close during the next quarter cycle connecting attenuator "B" into the circuit; during the third quarter cycle, switches "C" close, connecting the unknown; and all switches are open during the fourth quarter cycle, producing the zero base line. The resulting scope trace is shown in Figure 4e. The response curve is displayed with a zero base line and two reference lines. The position of the reference lines on the trace corresponds to the attenuation of two separate attenuators.

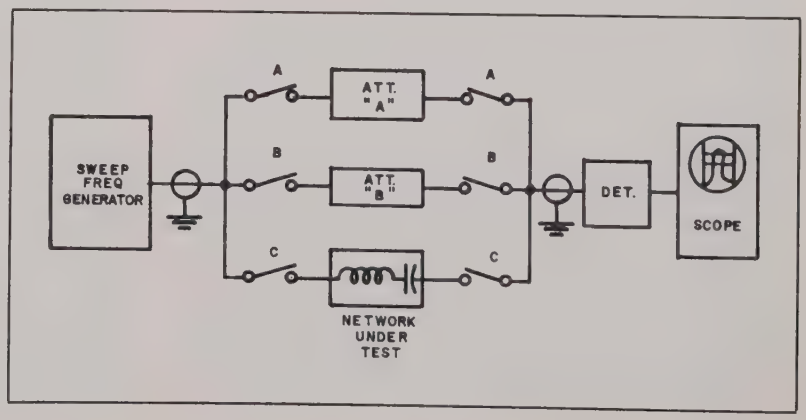


Fig. 4a
Operation of
3-Way Comparator
Model TC-3

This three-way display allows, for example, setting up a production test position with high and low limits of gain, loss, or impedance, keyed into the scope so that all calibrations are independent of generator output or scope drift.

Figure 5 shows a block diagram of the transistor driver employed to obtain this operation. With the input switch in INT position, the synchronizing input is obtained from a 60-cycle AC connection to the power supply through a 360° phase shifter. With the switch on EXT, an external synchronizing input may be used at any frequency between zero and several hundred cycles. A source capable of supplying six volts peak-to-peak into 20,000 ohms is required. Regardless of the input wave form, the switches will operate about one millisecond after each zero volt cross-over.

The internal or external synchronizing signal is shaped by a Schmitt trigger and clipper and the resulting square wave, which has the same frequency as the synchronizing source, is applied to a two-to-one counter generating a square wave at ½ that frequency. Both square wave voltages are applied to a diode AND gate, which controls the input to three driver transistors so that these are keyed in the sequence required to drive the relay coils. A function switch selects one of three switching mode connections to the driver transistors so that relay switching can be performed either: manually, or by three-way automatic switching, or by two-way automatic switching giving operation similar to the FD-30.

Three reed relays are used in each of two heads, mounted coaxially to obtain the best possible impedance match up to 1200 MC. The impedance match, when the switch is terminated in a precise 50 ohms, is held to a close tolerance as shown by the characteristic in Figure 6. The relay units were selected to give an unusually high degree of isolation when open, and low insertion loss when closed. The insertion loss, when the switch is closed, is less than 0.1 db from DC to 1200 MC. When open, the capacity is less than 0.085 pf giving an insertion loss which decreases with frequency, as shown in Figure 7.

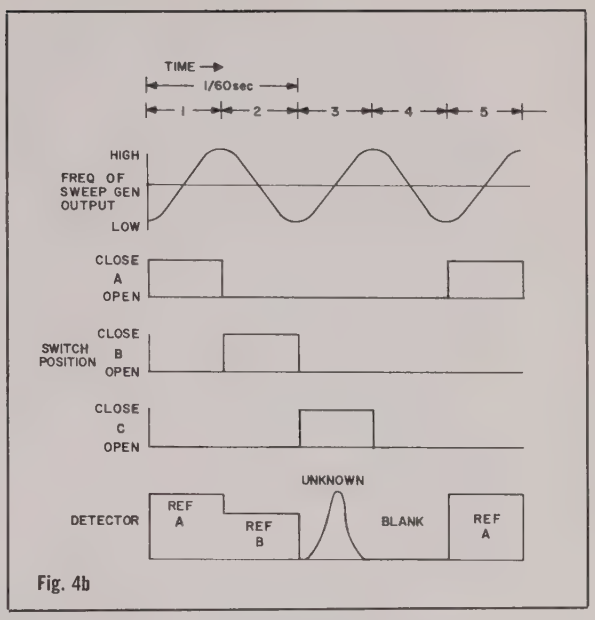


Fig. 4b

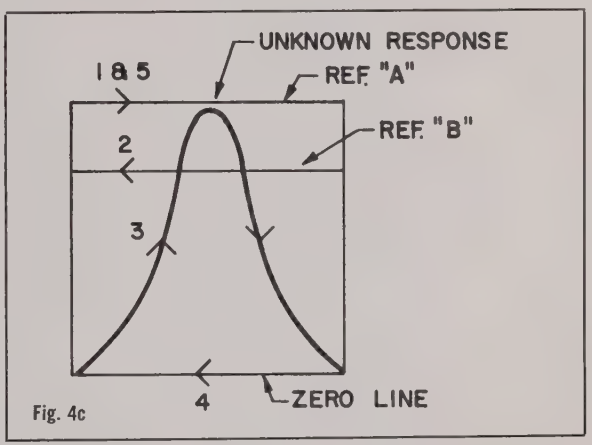


Fig. 4c

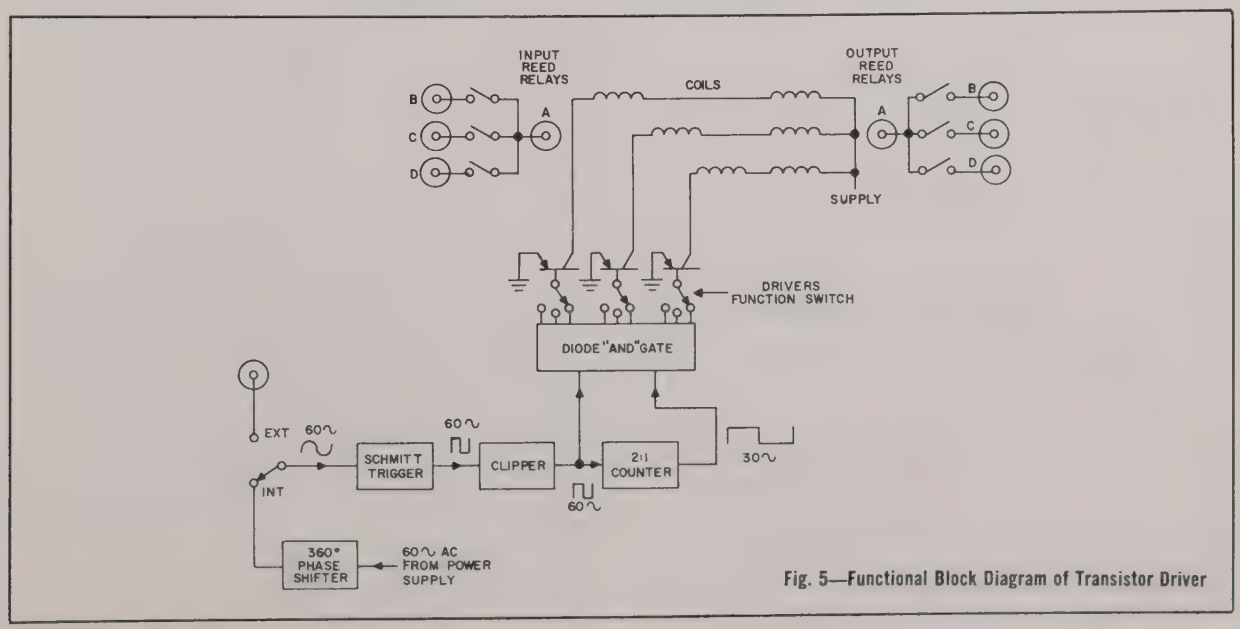


Fig. 5—Functional Block Diagram of Transistor Driver

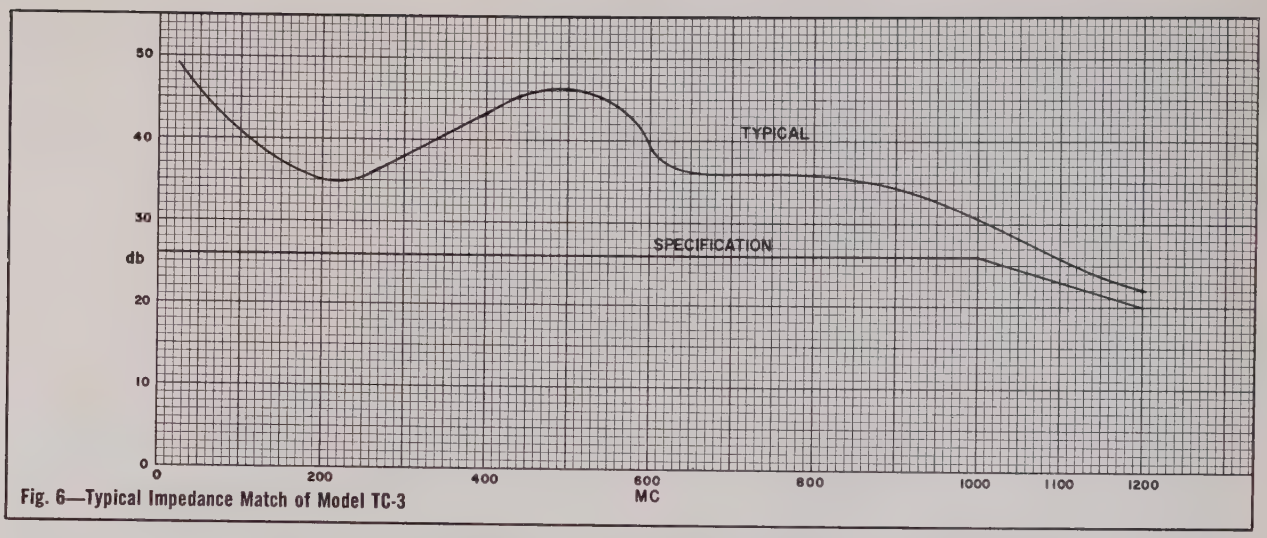


Fig. 6—Typical Impedance Match of Model TC-3

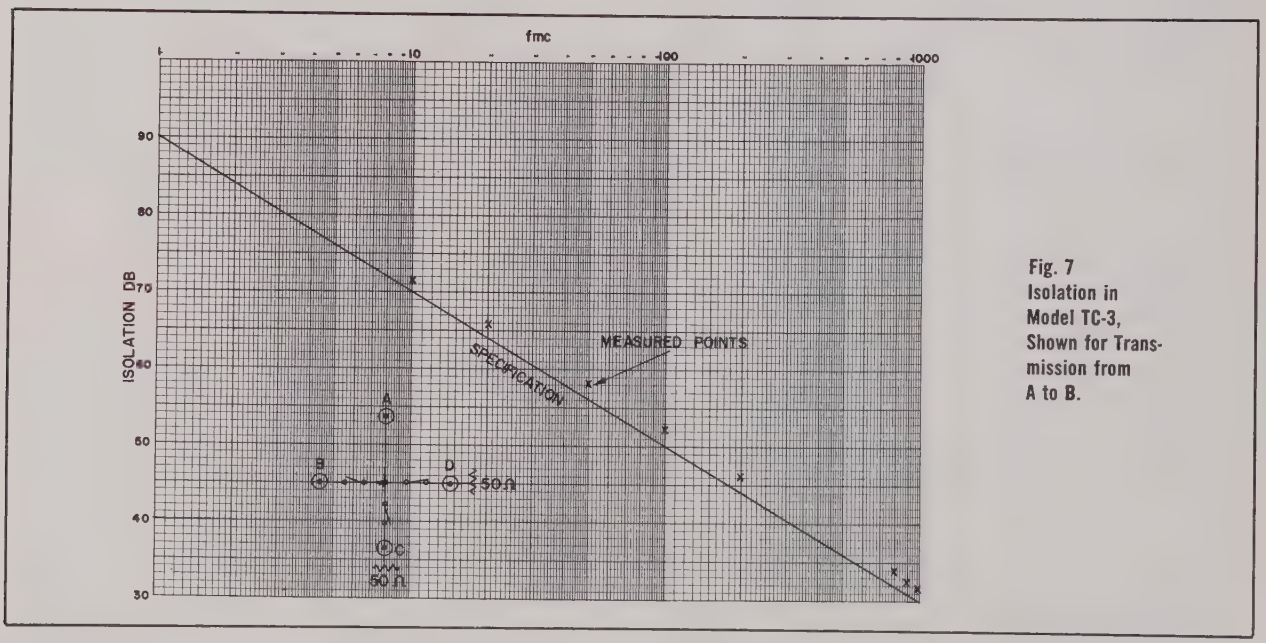


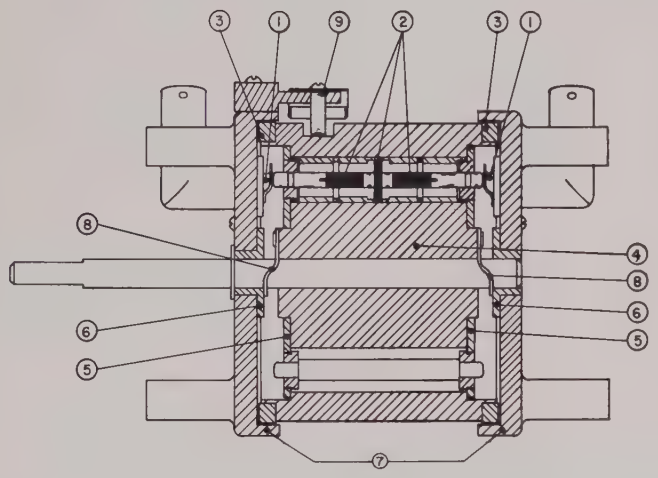
Fig. 7
Isolation in Model TC-3, Shown for Transmission from A to B.

MODEL ATV ATTENUATORS

The design selected for the new attenuators strikes a happy compromise between two extremes. Variable attenuators available today fall into two general categories. The best commercially available units are drum attenuators with an upper frequency limit of 2 to 4 GC. Several of the available types are pull-and-turn mechanisms which have optimum electrical performance at the expense of mechanical convenience. Prices are in the vicinity of $750 to $800 for a 10-position drum. All of the best units use disc and rod assemblies with either deposited carbon or film resistors.

At the other extreme there are many types of less expensive attenuators which cost less than $200. These generally use deposited or molded carbon resistors and various switching arrangements, either the progressively additive type (like our AV-50) or rotary switches. While these are excellent for low frequency applications and have good mechanical convenience, it is very difficult to maintain acceptable performance up to 1 GC with any arrangement where resistive elements are connected into the attenuator with wire leads.

Fig. 2 illustrates a rack-mounting bank of attenuators (Model ATV-109). A cross-section view of a turret is given in Figure 8. The highest quality deposited carbon rod and disc assemblies are used, mounted coaxially to take full advantage of their excellent match and attenuation capabilities. Spring contacts are employed between the attenuator element and the input and output connectors. While this construction does not provide as good a coaxial connection as might be obtained with a pull-and-turn arrangement, with careful design it has been made to give excellent performance to our top frequency objective of 1200 MC. The carefully-machined castings and well-designed detent ensure accurate and trouble-free operation. Complete RF shielding is assured by the use of RF gasketing material. The impedance match characteristics and attenuation accuracy of a typical 0 - 50 db drum are illustrated in Figures 9 and 10.



1. Coin silver contacts.
2. Disc and rod pad resistors for accuracy and long time-stability. Low VSWR assured with machined compensating sleeves.
3. Resilient rf gasketing for good ground contact and shielding.
4. Solid aluminum rotor, precision machined for accuracy and stability.
5. Important rf grounds carried by silver plated brass rings.
6. Brass bearing inserts for minimum wear.
7. Machined aluminum end castings to maintain alignment.
8. Coin silver inner contact grounding spring (necessary on some models only).
9. Positive spring-loaded ball bearing detent mechanism with faultless resolution.

Fig. 8—Cross-Section of Turret Attenuator

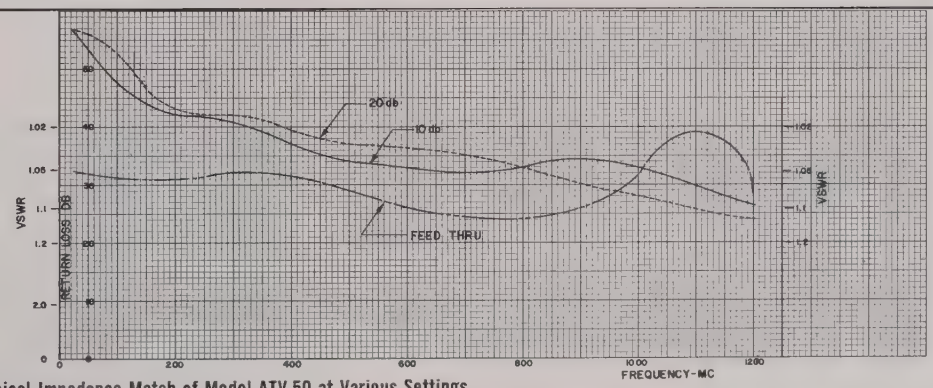


Fig. 9—Typical Impedance Match of Model ATV-50 at Various Settings

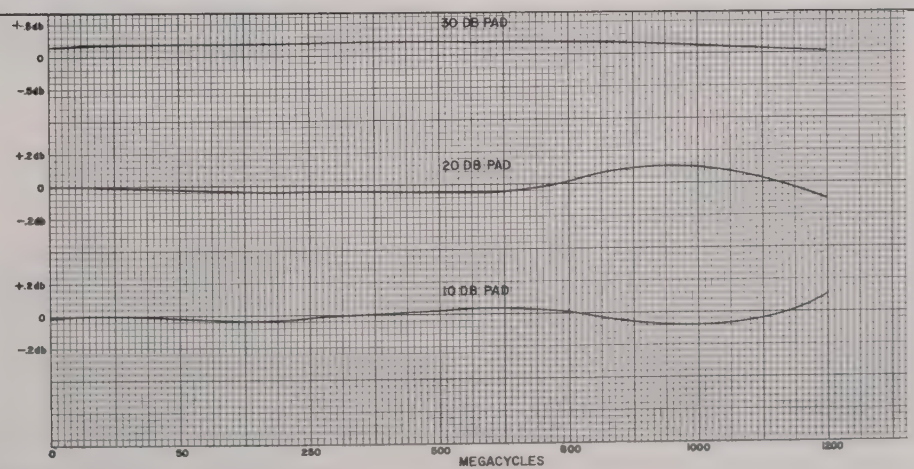


Fig. 10—Typical Attenuation Tolerance of Model ATV-50

Design of the attenuators and comparators was carried out under the direction of Henry Arbeiter. He also is responsible for electrical design of the attenuator, working with Eric Winston on mechanical design. John Austin designed the reed-relay heads on the TC-3, Frank Egenstafer is responsible for the transistor driver, and Earl Nickerson for the styling.

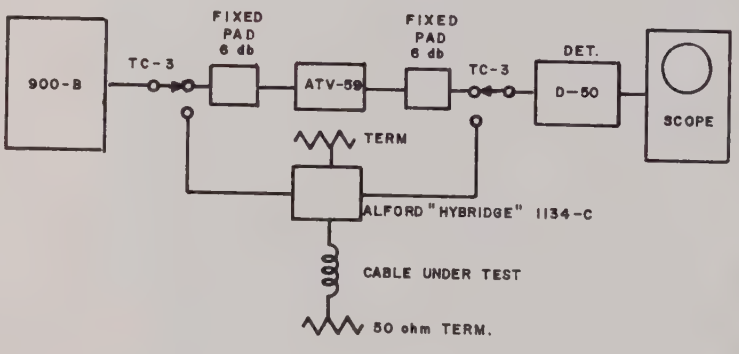
MODEL 890 WIDE-BAND SWEEPER

Model 890

Latest in Jerrold's proven group of sweep frequency generators is the new, moderately priced Model 890. This sweeper, although somewhat less sophisticated in its design than the Model 900-A, nevertheless incorporates all of that instrument's basic circuitry necessary to permit a highly economical production line test set-up.

Together with the new comparator and attenuators, the 890 will afford fast, repetitive measurements for simultaneous low and high limit comparison. Tests can be made up into the UHF range of 1100 MC, at a top center frequency setting of 1000 MC with a sweep width variable from 100 KC min. to ± 10% of center frequency.

TYPICAL APPLICATIONS IN MEASUREMENTS BY COMPARISON



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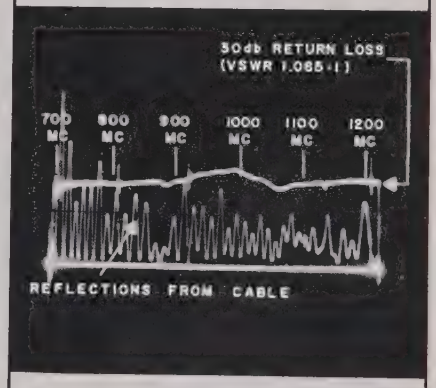
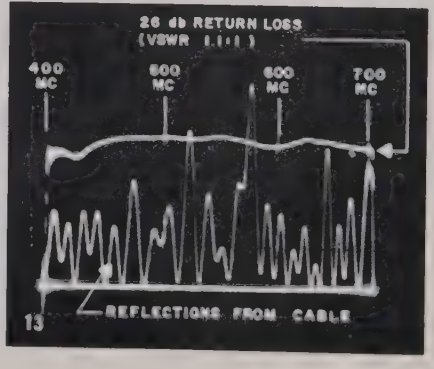


12

Application of Models TC-3 and ATV-50 in testing the impedance characteristics of coaxial cables. The test involved approximately 50 ft. of type RG-8A/U flexible coaxial cable, carefully terminated at the other end. Photo 12 illustrates the trace for measurement between 700 and 1200 MC; photo 13 illustrates the trace for measurement between 400 and 700 MC. The impedance variations on each trace are due to mechanical irregularities in the tested cable.

Note that there are decidedly less irregularities in photo 12, where the cable would meet a specification of 30 db (better than most cables).



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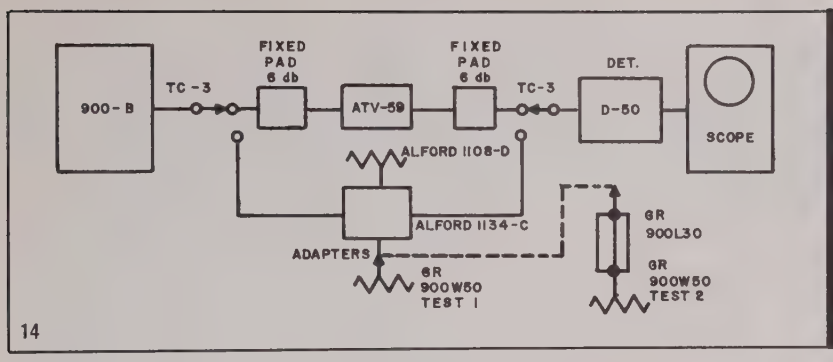
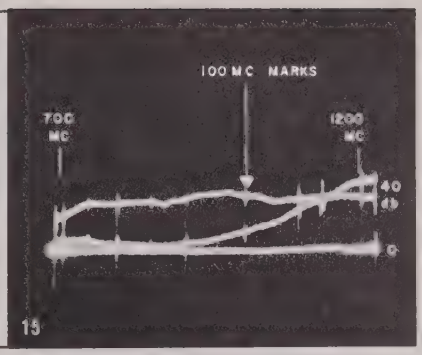




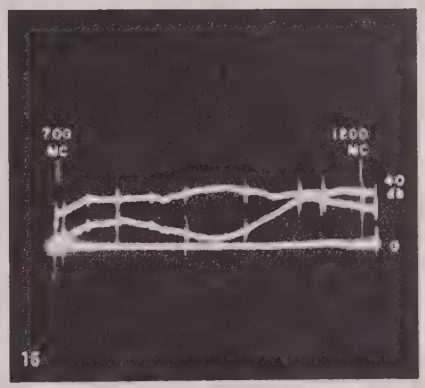
Application of Models TC-3 and ATV-50 in testing the sensitivity of Model 900-B and for checking the accuracy of the bridge test set-up.
Two tests were made:

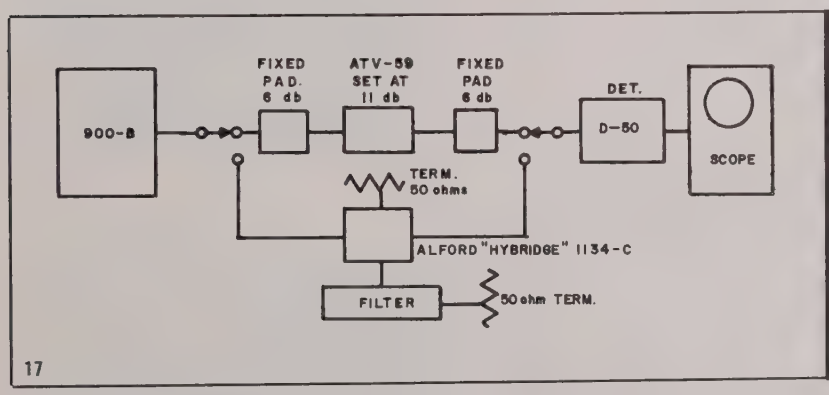
1. connecting a General Radio type 900W50 precision 50-ohm termination through two adapters, first a General Radio type 900Q874, followed by a General Radio type 874QNP, to the bridge.

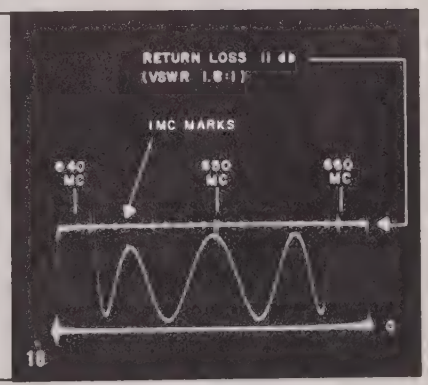
   Photo 15 shows that the bridge measured somewhat better than 40 db match (VSWR 1.02:1) up to 1100 MC.

2. connecting a 30 cm length of precise 50-ohm air dielectric line, General Radio type 900L30, between the termination and the 900Q874 adapter.

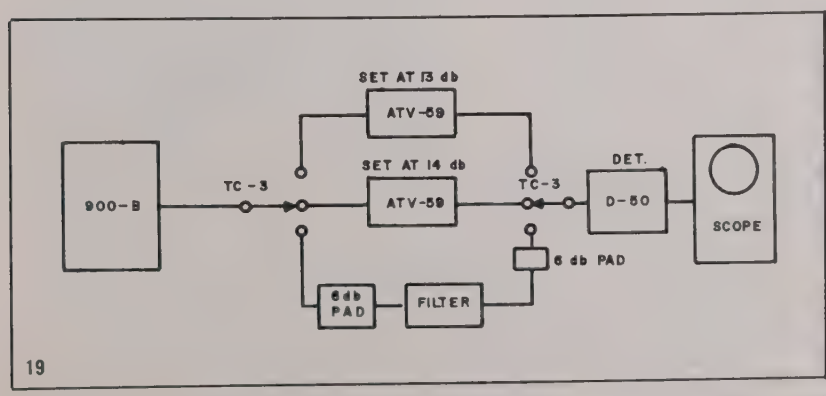
   If the characteristic impedance of the air line were exactly equal to that of the termination, the impedance would be unchanged in the second test. As seen in photo 16, there was a slight change, but the resulting impedance was still seen by the bridge as better than a 40 db match.

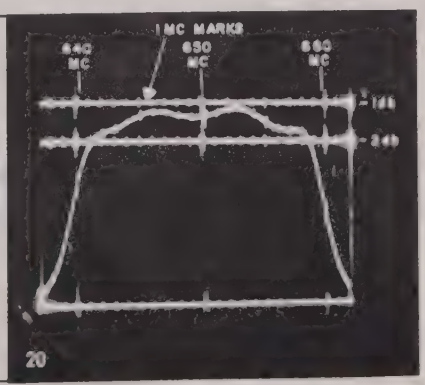




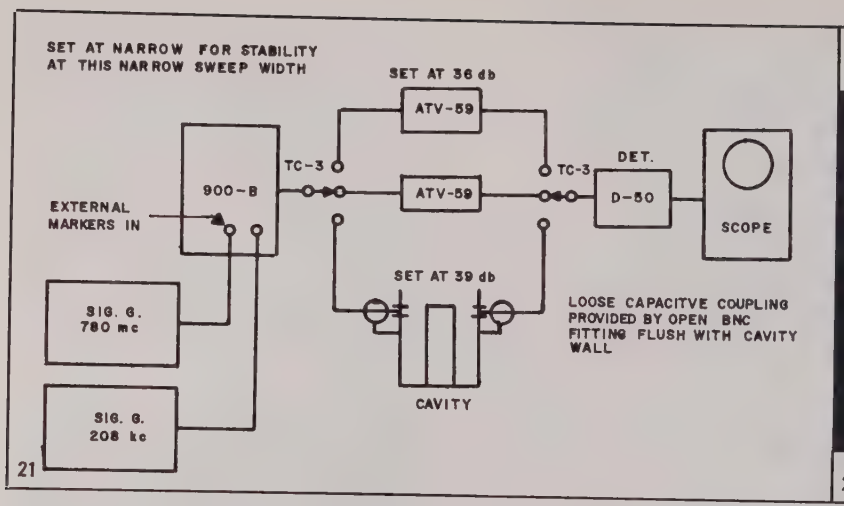


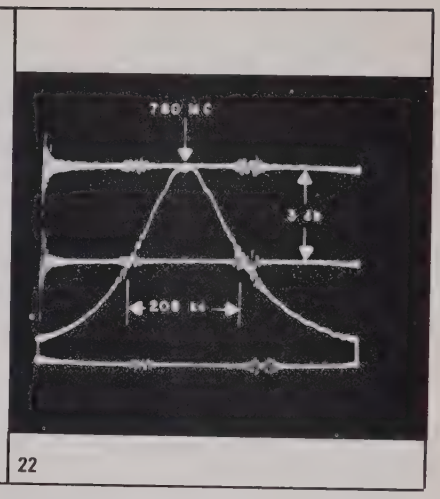
Application of Models TC-3 and ATV-50 to impedance match measurement on a four-pole band-pass filter.



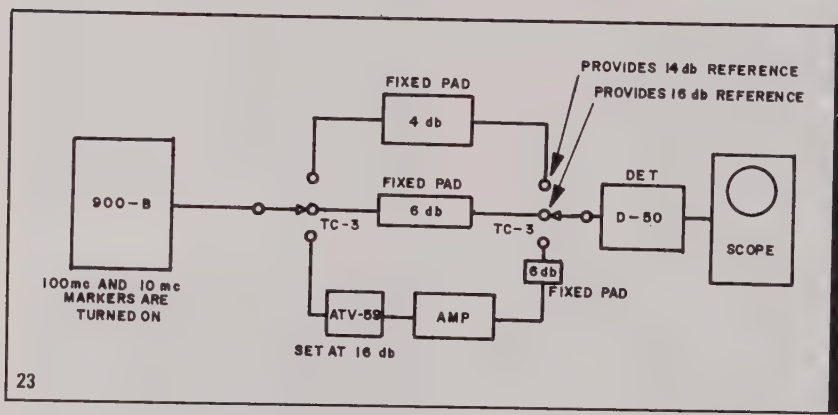


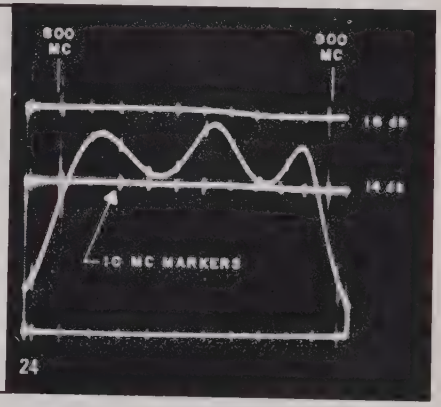
Application of Models TC-3 and ATV-50 to measurement of insertion loss of filter the match of which was shown in photo 18.



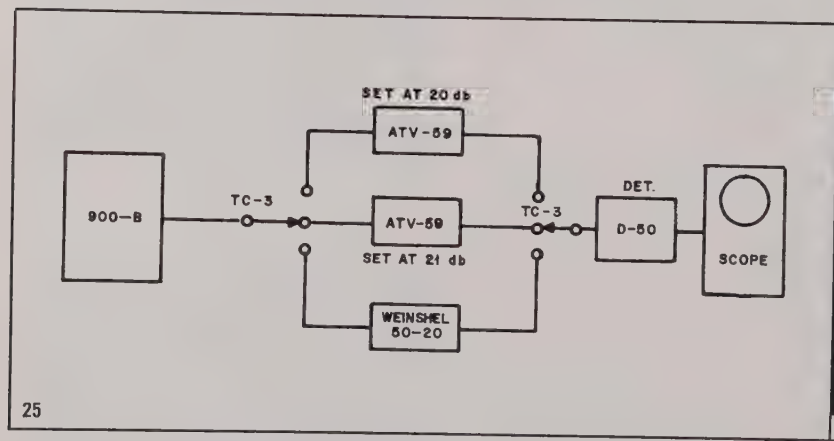


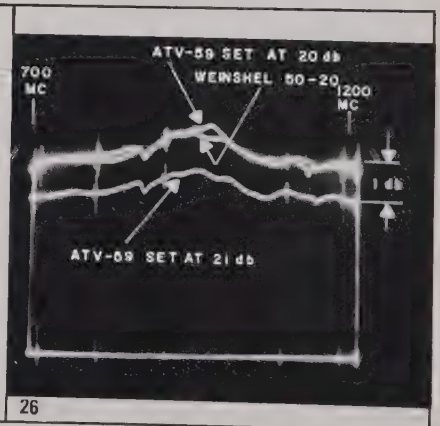
Application of Models TC-3 and ATV-50 to measurement of unloaded bandwidth and thus Q of 780 mc resonant cavity. Resulting $Q = \frac{fo}{\Delta f} = \frac{780}{0.208} = 3750$





Application of Models TC-3 and ATV-50 to measurement of gain of a transistorized UHF amplifier. Gain limits are 16 and 14 db; design bandwidth was 806 to 890 MC.





Comparison of Model ATV-50 with a Weinshel attenuator pad which had been calibrated at 20.0 db at 1000 MC. Differences between the two curves are about 0.1 db maximum, some of which is due to short cable jumpers.

NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila., Pa. 19132. Phone: BAldwin 6-3456 — Ext. 215.

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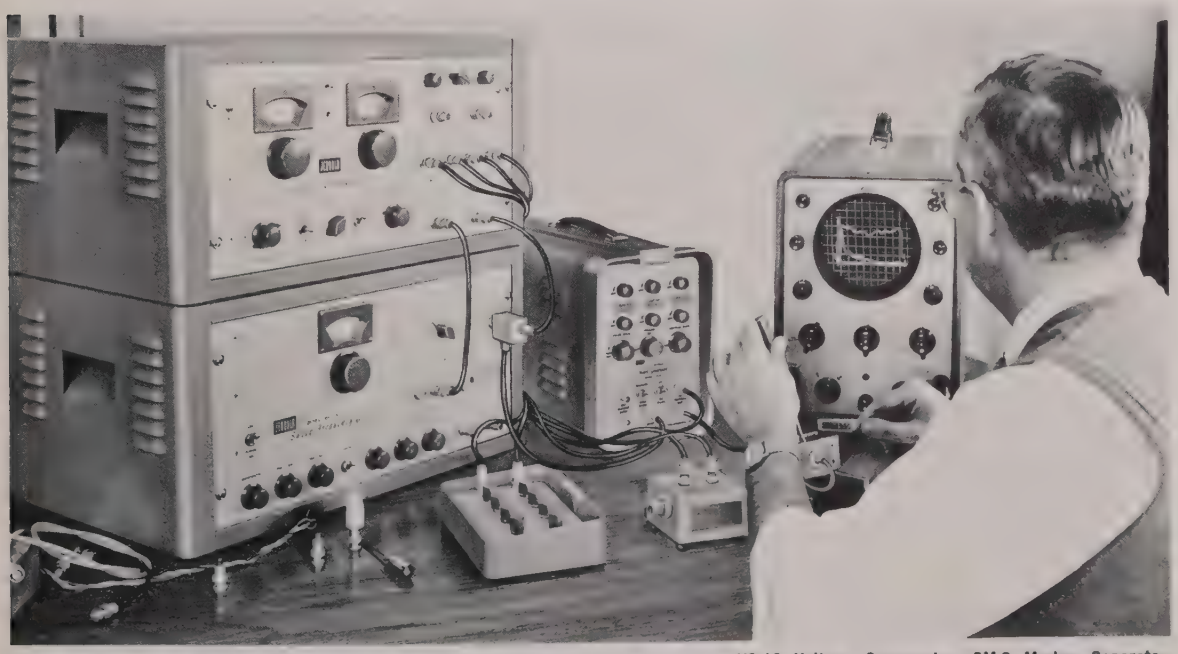
A Bridge Method of Sweep Frequency Impedance Measurement

by KEN SIMONS, Chief Engineer, Jerrold Electronics Corp.

The following is an extract of a speech given by Mr. Simons at the First Annual Eastern Engineering Seminar, May 5, 1960, Hotel Bradford, Boston, Massachusetts.

SUMMARY:

A matched bridge circuit is described providing oscilloscopic display and measurement of return loss vs. frequency. Units have been constructed allowing measurement of 50 db return loss over a 50 to 1 frequency range (4-200 MC). The technique is useful from low audio frequencies up to several hundred megacycles.



A typical test set-up utilizing Jerrold Models—900-A Sweep Frequency Generator, VC-12 Voltage Comparator, CM-6 Marker Generator, AV-75 Variable Attenuator and KSB-75 ohm Bridge; to show an optimum termination, 4 mcs to 100 mcs, with a 50 db reference.

FUNDAMENTALS:

Since high frequency measurements frequently require unavoidable lengths of transmission line, high frequency measuring techniques usually involve an application of transmission line principles. The bridge circuit to be described is no exception. Understanding its operation requires a clear picture of the concepts of reflection coefficient and return loss.

Consider an AC generator with internal impedance Ro and a voltage E as illustrated in Figure 1 (a). When the output

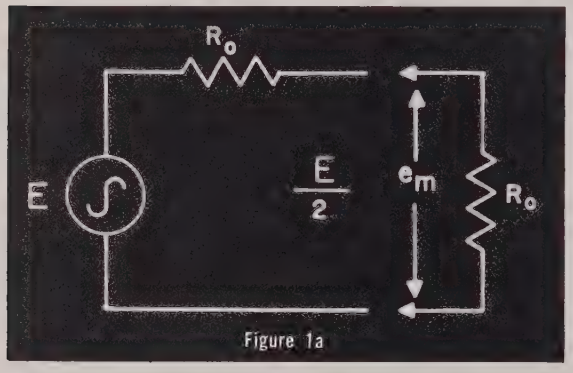


Figure 1a

of this generator is connected to a load having an impedance equal to Ro, the output voltage is $\frac{E}{2}$. In the particular case where this load is the input impedance of a section of transmission line having a characteristic impedance Ro, it is convenient to consider the input voltage as consisting of two components—a main voltage wave (e_m) which travels down the line away from the generator, and a reflected wave (e_r) which travels back from the far end towards the generator. When the transmission line section is terminated in Ro, the reflected wave is zero and the input voltage is simply e_m Figure 1 (b).

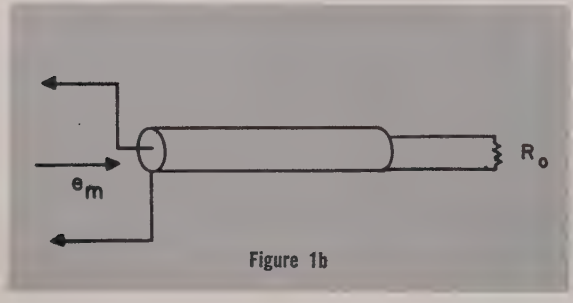


Figure 1b

This terminology can be applied even when there is no transmission line in the circuit. It may correctly be said in reference to Figure 1 (a) that the voltage across the load Ro is e_m.

In the case where the generator feeds a section of transmission line terminated in some impedance other than Ro, the input voltage can be considered to be the sum of e_m and e_r (Figure 1c). Similarly the voltage across any load

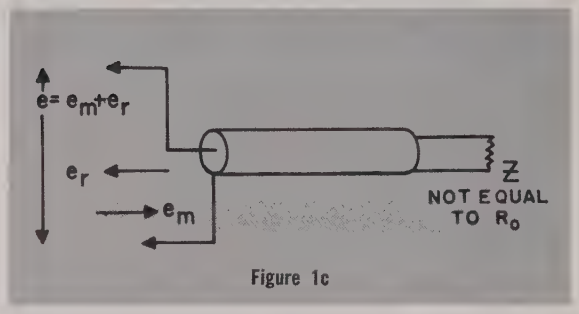


Figure 1c

connected to a generator can be considered to be the sum of a main wave and a reflected wave (Figure 1d).

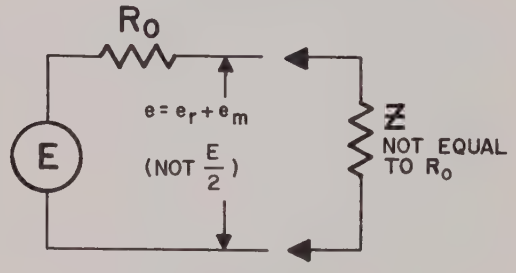
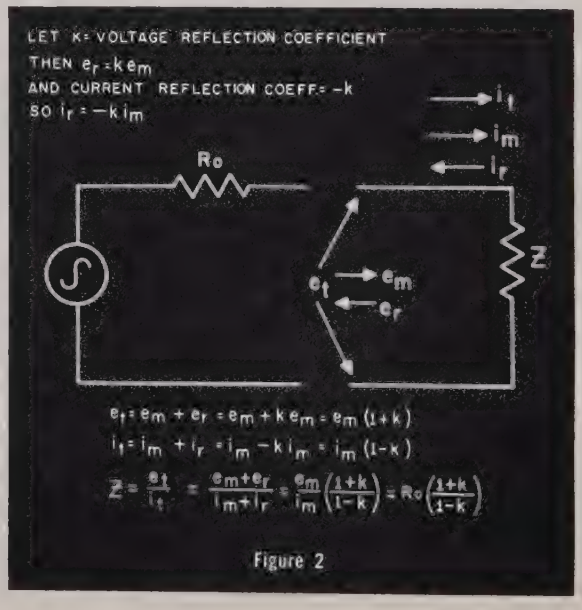


Figure 1d

A convenient way of describing the relation between the main and reflected waves is to use the reflection coefficient. Figure 2 illustrates a generator feeding a load having any impedance "Z". The voltage and current across this load


LET K= VOLTAGE REFLECTION COEFFICIENT
THEN er = k em
AND CURRENT REFLECTION COEFF= -k
SO ir = -k im
it
im
ir
Ro
Z
et
em
er
et = em + er = em + k em = em (1+k)
it = im + ir = im - k im = im (1-k)
Z = et/it = (em+er)/(im+ir) = (em/im)((1+k)/(1-k)) = Ro((1+k)/(1-k))
Figure 2

are expressed in terms of their main and reflected components. The quantity k is defined as the ratio of the reflected voltage e_r to the main voltage e_m. The current reflection coefficient is numerically equal to the voltage reflection coefficient and has the opposite sign, so that the relationships indicated in Figure 2 hold true. A significant point is that the impedance of the load can be fully described by the internal impedance of the generator and the voltage reflection coefficient. The relation is:

$$Z = Ro \frac{1 + k}{1 - k}$$

Note that k is a complex number involving the relation between not only the magnitudes of the main wave and reflected wave, but also their phase angles.

A further convenience is achieved by expressing the reflection coefficient logarithmically. The "Return Loss" is defined as 20 $\log_{10}\frac{1}{k}$. Thus, it is the ratio of e_r to e_m expressed in decibels. The interrelation between Return Loss, Reflection Coefficient and VSWR is indicated on the Nomograph of Figure 3.

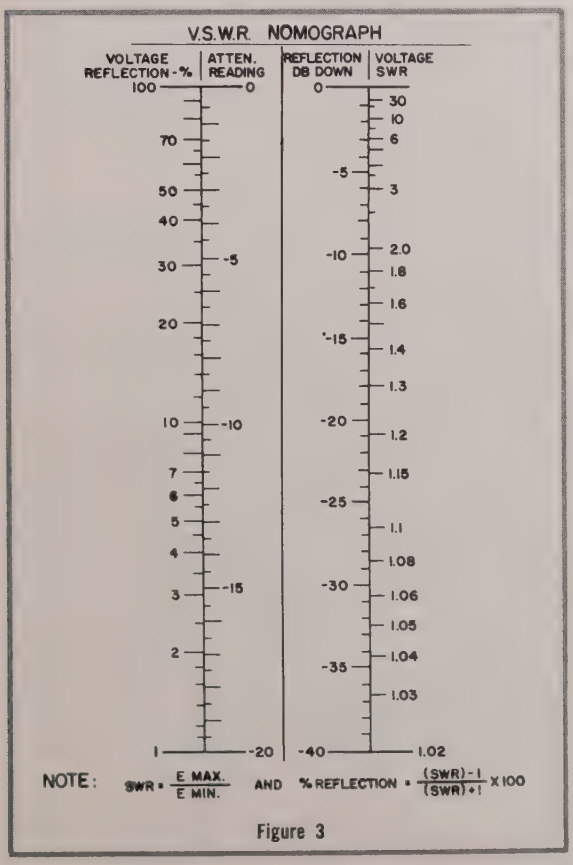


Figure 3

THE BRIDGE PRINCIPLE:

A convenient way of measuring an unknown impedance would be to separate the reflected wave from the main wave and to measure its relative magnitude and phase angle. This is commonly done in high-frequency and microwave testing by the use of a directional coupler. Another approach is the use of a bridge.

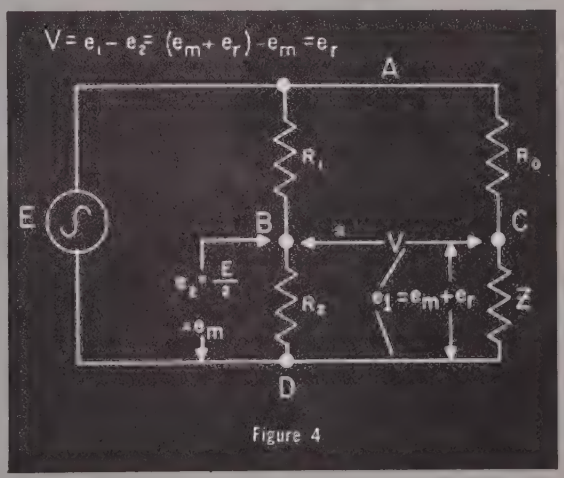


Figure 4

Figure 4 illustrates how a bridge can provide a voltage equal to the reflected wave. A generator having zero internal impedance feeds a load Z through a resistor Ro. The voltage e_r across Z is equal by definition to e_m plus e_r. If a second divider consisting of two equal resistances R_1 and R_2 is connected across the voltage source, the voltage e_2 across the lower resistor will be equal to $\frac{E}{2}$ which is the same as e_m. If the voltage between the center points of the two dividers is measured, it is found to be the difference between e_m plus e_r and e_m, which is e_r. Thus, the output voltage "V" of this network is identical to the reflected voltage component from the impedance Z.

THE TERMINATED BRIDGE:

Unfortunately, this simple circuit involves the use of a voltage source having zero internal impedance and a voltage measuring device having infinite impedance which measures the difference of voltage between two points, neither of them grounded. As a practical matter, these conditions are difficult to approach.

Fortunately, an alternative configuration is available. The terminated bridge illustrated in Figure 5 has several characteristics which make it especially suitable for this application. When all six resistances are equal, the impedance seen looking into the network at the terminals left after disconnecting any one of the resistances is always equal to Ro. Thus, for example, in Figure 5 the impedance seen between A and D with #1 resistor removed is equal to Ro. When the resistors are equal to the characteristic impedance of a particular coaxial cable, the bridge has matched input and output impedances.

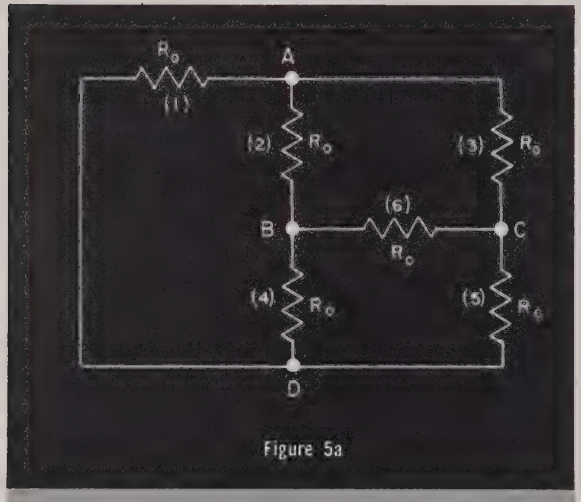


Figure 5a

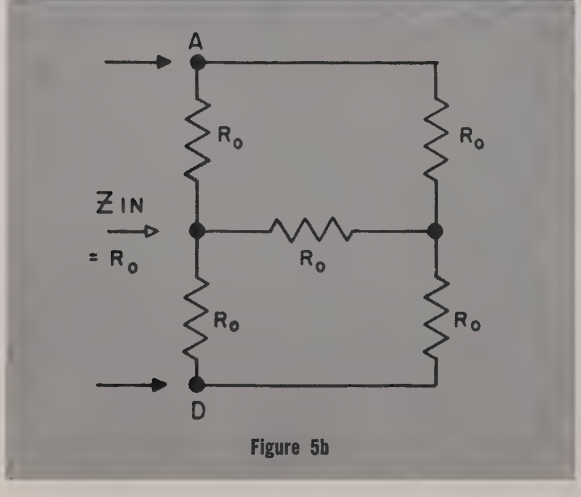
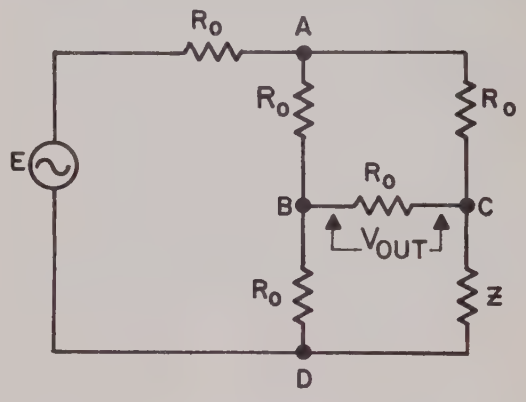


Figure 5b

The terminated bridge has the further advantage that it provides an output voltage precisely equal in magnitude and phase to a constant times the reflection coefficient of the unknown. This is illustrated in Figure 6 which states



$$\text{TRANSMISSION LOSS IS } 20 \log_{10} \left(\frac{E}{2\, V_{OUT}} \right)$$

$$= 20 \log_{10} 4 \times \frac{Z+R_o}{Z-R_o} = \text{RETURN LOSS OF } Z+12\,\text{db}$$

Figure 6

that the transmission loss from the input terminals of the bridge, to Vout is equal to the return loss of Z (referred to Ro) plus 12 db. This relation is fully derived in Appendix I. It may be understood better by referring to Figures 7 and 8. Figure 7 shows that the input voltage to a matched load connected between C and D (e_m) is equal to the bridge input Vin reduced by 6 db.

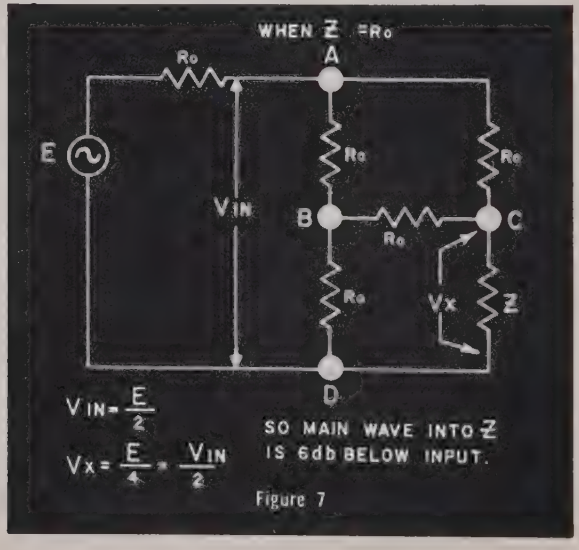


Figure 7

In Figure 8 the bridge is redrawn to show the voltage division experienced by the return wave from Z. It comes back into terminals C and D and is attenuated 6 db before coming out of terminals B and C.

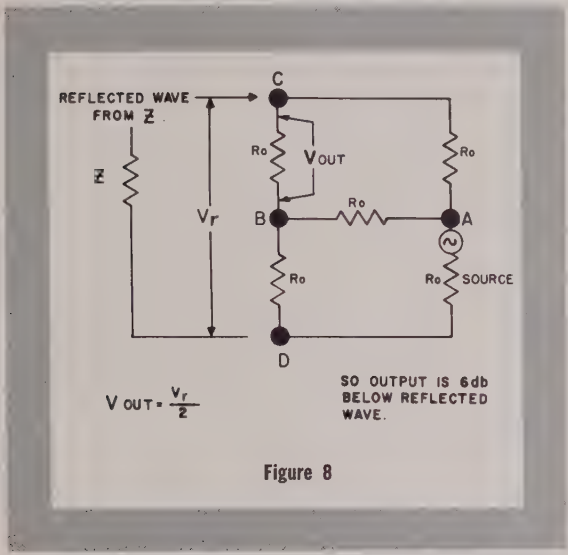


Figure 8

As compared with the simple bridge circuit illustrated in Figure 4, the matched bridge has the advantage that the input and output impedances match connecting coaxial cables so that wide band frequency performance can be obtained. The problem of providing a differential voltage measurement between two ungrounded terminals is still present. To build a matched bridge practically, it is necessary to provide a transformer having balanced input terminals as illustrated in Figure 9. It has been found

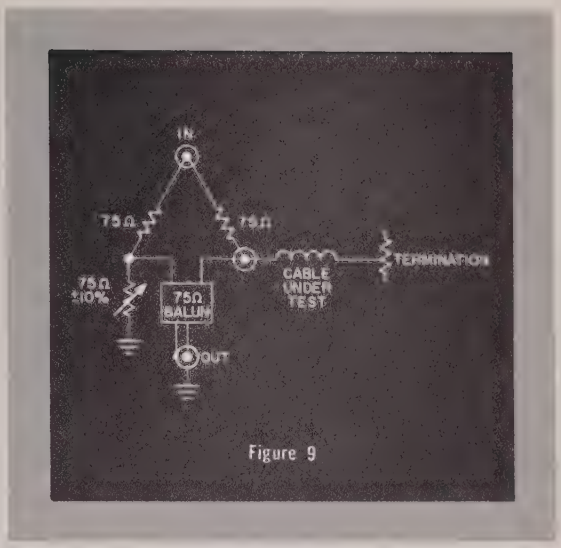


Figure 9

possible to build such a transformer with a very wide range of frequency response. The resulting bridge is conveniently connected to provide a calibrated sweep display as illustrated in Figure 10. A high-speed coaxial switch (Jerrold Model FD-30) is connected so that in the "up" position it connects the output of the sweep through a 12 db pad and a standard variable attenuator to the input of a wide band amplifier. The amplifier is followed by a detector and scope and displays a curve representing loss vs. frequency. In the "down" position the switches insert the bridge in place of the attenuators. Since the loss of the bridge is equal to the return loss of the unknown plus 12 db, the scope will show a plot of return loss vs. frequency with a reference line corresponding to the setting of the standard variable attenuator.

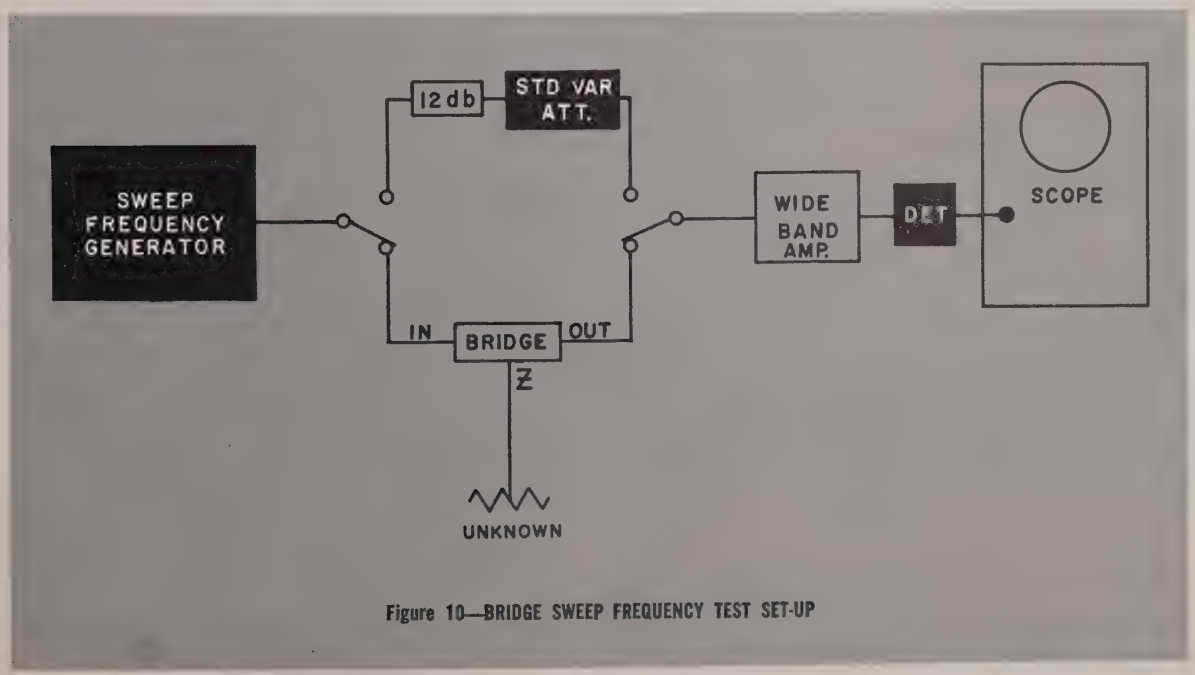


Figure 10—BRIDGE SWEEP FREQUENCY TEST SET-UP

EXAMPLES OF USE

The performance of a particular bridge is illustrated in Figures 11 to 16 inclusive. This bridge was designed for a characteristic impedance of 75 ohms and the test circuit covered a frequency range from 4 to 100 MC. Figure 11 illustrates the relation between the reference line and the bridge response with the unknown terminal of the bridge open-circuited. The bridge indicates 0 db return loss within about ½ db from 4 to 100 MC.

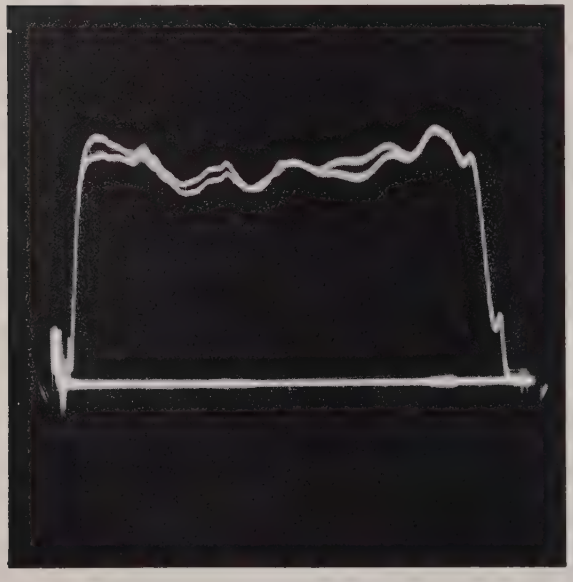


Figure 11—BRIDGE RESPONSE
4-100 mc, Open Circuit

Figure 12 shows the same situation with the unknown terminal short-circuited, again indicating 0 db return loss with approximately ½ db error.

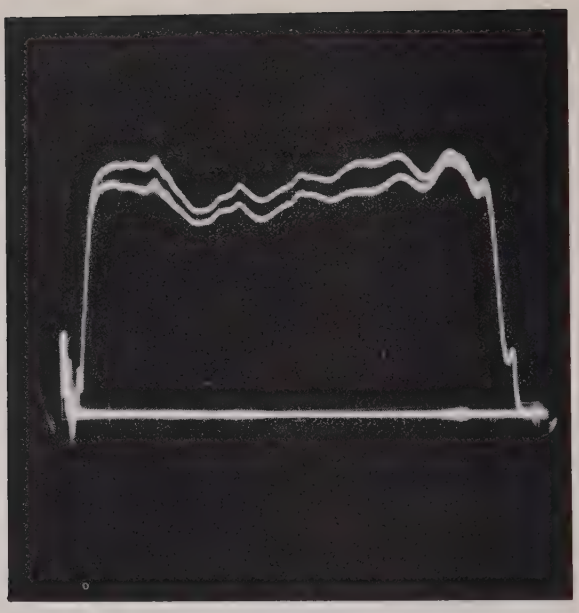


Figure 12—BRIDGE RESPONSE
4-100 mc, Short Circuit

To obtain Figure 13, the X terminal of the bridge was terminated as accurately as possible. In this figure the reference line represents a return loss of 50 db and it is seen that the loss through the bridge indicates a return loss substantially better than 50 db across this entire frequency range. A 50 db return loss corresponds to a VSWR of 1.06 so that it is apparent that this technique is capable of precision comparable with the best that can be obtained with other high frequency measuring techniques.

Figure 13—BRIDGE RESPONSE
4-100 mc, Optimum Termination, 50 db Reference

Although this method is extremely **precise**, i.e., it can compare an unknown with a standard and show deviations of considerably less than 1%, its **accuracy** is entirely dependent on the accuracy of the standard impedance against which the bridge is tested. Where suitable standards are available, its accuracy can be nearly as good as its precision.

Figure 14 illustrates the agreement between the reference line and the bridge loss when the bridge was first adjusted for balance with a 75 ohm resistor. A resistor was then substituted whose DC resistance was accurately adjusted to 70 ohms. A 70 ohm resistor in a 75 ohm system gives a VSWR of 1.07 or a return loss of 29.2 db. The reference line on this trace was 30 db.

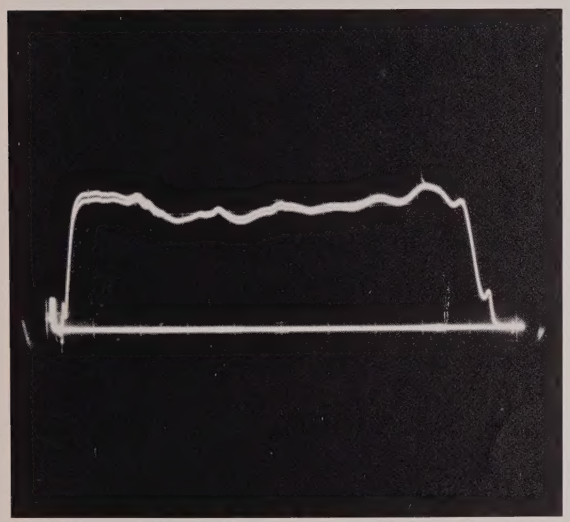


Figure 14—BRIDGE RESPONSE
4-100 mc, 70 ohm Termination, 30 db Reference

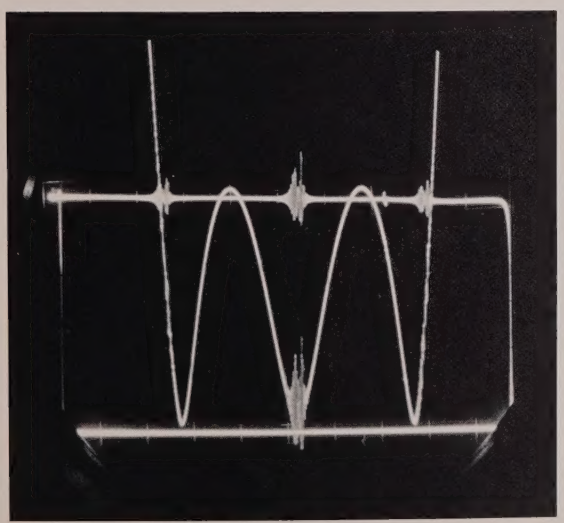


Figure 15 —BRIDGE RESPONSE
Markers 100 mc ±1.5 mc, 10 db Reference

The delay line method of sweep-frequency impedance measurement (as described in Technical Newsletter #2) provides a convenient method of measuring high frequency impedances with a minimum of equipment. Practical characteristics of delay lines limit the usefulness of this method to relatively wide frequency bands. The bridge technique has no such limitation. It is equally useful for extremely narrow-band measurements or extremely wide-band measurements.

Figure 15 shows the return loss measurement of a triple-tuned filter having a bandwidth of 1.5 MC and a maximum return loss in the pass band of approximately 10 db. Notice the simplicity of this presentation and the ease with which an operator could adjust such a filter to meet given frequency and return loss specifications.

Figure 16 illustrates an extremely wide band measurement. This shows the return loss of a piece of high-grade coaxial cable over the frequency range 4 to 100 MC with a 40 db reference. The bridge provides a way of testing this complicated function with a high degree of speed and accuracy.

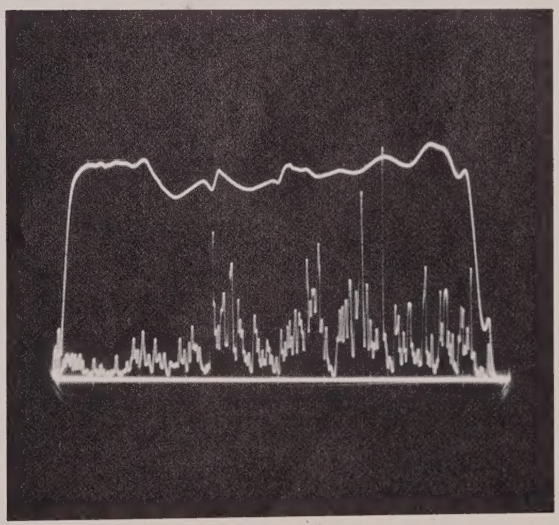
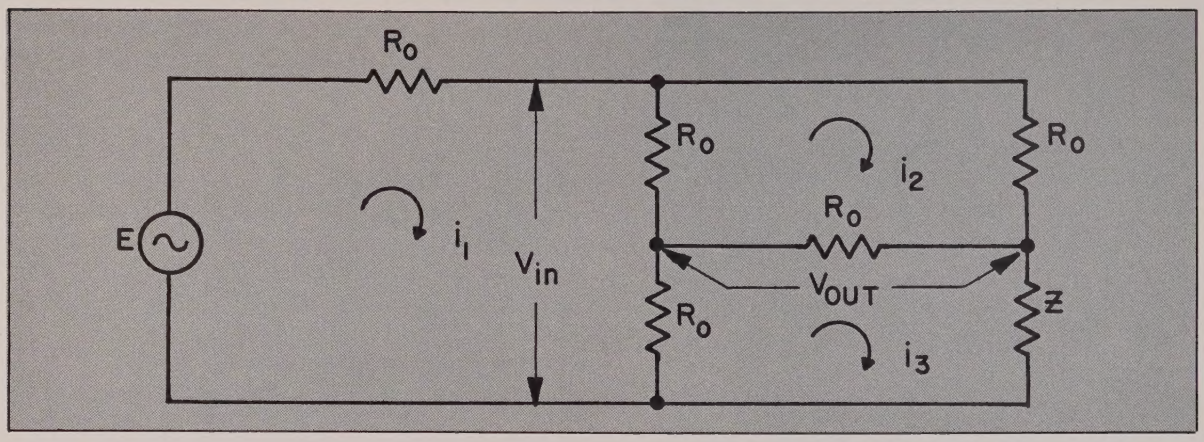


Figure 16—BRIDGE RESPONSE
4-100 mc, Input of 1000′ Piece High Grade Coax., 40 db Reference

The sweep frequency technique has for many years provided a convenient way of observing the frequency variation of gain or loss. More recently it has become possible to measure these quantities simultaneously with the sweep presentation. (See Technical Newsletter #1.) The wide-band return-loss bridge now provides the same degree of convenience in viewing and measuring the frequency variation of return loss and opens a new area of speed and convenience in Laboratory and Production testing of high frequency networks.

APPENDIX



① $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots E = 3R_0 i_1 - R_0 i_2 - R_0 i_3$

② $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots 0 = -R_0 i_1 + 3R_0 i_2 - R_0 i_3$

③ $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots 0 = -R_0 i_1 - R_0 i_2 + (Z + 2R_0) i_3$

④ ② − ③ $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots 0 = 4R_0 i_2 - (Z + 3R_0) i_3$

⑤ ① + (3 × ②) $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots E = 8R_0 i_2 - 4R_0 i_3$

⑥ Solving ④ for i_2 $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots i_2 = \dfrac{Z + 3R_0}{4R_0}\, i_3$

⑦ Substituting ⑥ in ⑤ $\ldots\ldots\ldots E = 8R_0\left(\dfrac{Z + R_0}{4R_0}\right) i_3 - 4R_0 i_3 = 2(Z + R_0) i_3$

$$\text{hence } i_3 = \frac{E}{2(Z + R_0)}$$

⑧ Substituting ⑦ in ⑥ $\ldots\ldots i_2 = \left(\dfrac{Z + 3R_0}{4R_0}\right)\left(\dfrac{E}{2(Z + R_0)}\right) = \left(\dfrac{E}{8R_0}\right)\left(\dfrac{Z + 3R_0}{Z + R_0}\right)$

⑨ $\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots V_{OUT} = R_0 (i_2 - i_3) = R_0\left[\left(\dfrac{E}{8R_0}\right)\left(\dfrac{Z + 3R_0}{Z + R_0}\right) - \dfrac{E}{2(Z + R_0)}\right];$

$$\text{hence } V_{OUT} = \frac{E}{8}\left(\frac{Z + 3R_0 - 4R_0}{Z + R_0}\right)$$

$$\text{and } \boxed{V_{OUT} = \frac{E}{8}\left(\frac{Z - R_0}{Z + R_0}\right)}$$

$$\text{Transmission Loss} = 20 \log_{10}\left(\frac{E}{2 V_{OUT}}\right) = 20 \log_{10}\left[\frac{E8\,(Z + R_0)}{2E\,(Z - R_0)}\right] =$$

$$= 20 \log_{10} 4\left(\frac{Z + R_0}{Z - R_0}\right) = \text{Return Loss of } Z + 12\,\text{db}$$

NOTE— Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtanied by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila., Pa. 19132. Phone: BAldwin 6-3456 — Ext. 215.

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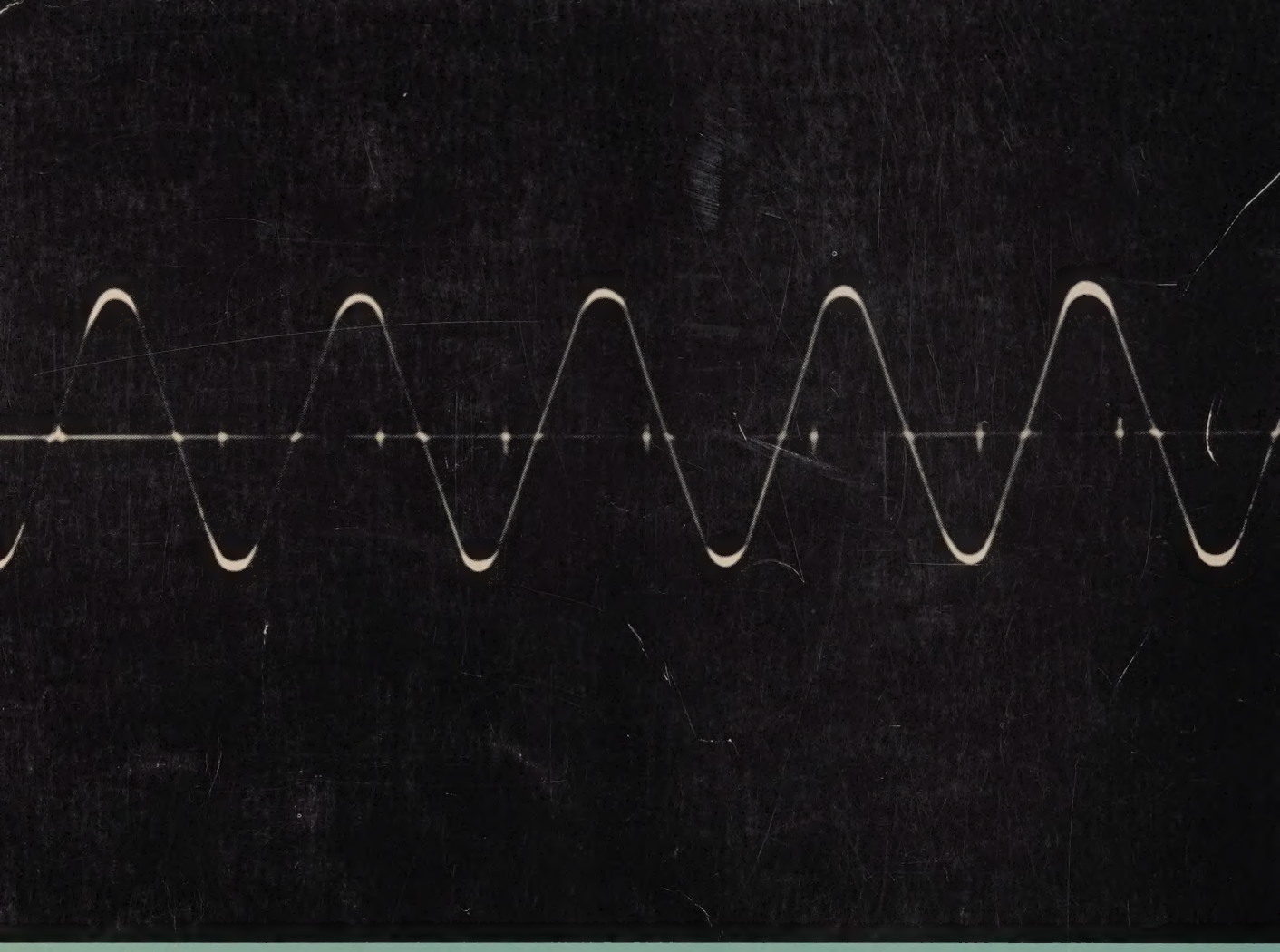

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